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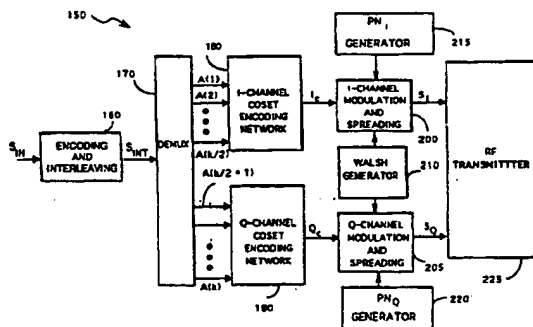
(57) Abstract

An improved system and method for communicating information at variable data rates over in-phase (I) and quadrature phase (Q) communication channels in a CDMA spread spectrum communication system is disclosed herein. In an exemplary implementation an input information signal is transmitted over either an I or Q communication channel using a direct sequence spread spectrum communication signal. The information signal is initially divided into first and second subsignals, which are respectively provided to first and second coset-encoding networks. The first coset-encoding combines the first subsignal with a first coset code, while the second coset-encoding network combines the second subsignal with a second cosed code orthogonal to the first cosed code. In this way the first and second coset-encoding networks are operative to produce first and second coset-encoding signals, respectively. A composite coset-encoded signal formed from the first and second coset-encoded signals is then modulated by an orthogonal function signal to provide a first modulated signal. In-phase pseudorandom noise (PN$_I$) and quadrature phase pseudorandom noise (PN$_Q$) signals of predetermined PN codes are used for spreading the first modulated signal for transmission to a receiver over either the I or Q communication channel, respectively. The receiver is operative to produce an estimate of the input information signal on the basis of the modulated carrier signal received over either the I or Q communication channel.

Variable Rate Signal Transmission in a Spread Spectrum Communication
System Using Coset Coding

BACKGROUND OF THE INVENTION

I.   Field of the Invention

The present invention relates to communication systems utilizing spread
spectrum signals, and, more particularly, to a novel and improved method and
5   apparatus for communicating information in a spread spectrum communication system.

II.   Description of the Related Art

Communication systems have been developed to allow transmission of
information signals from a source location to a physically distinct user destination.
10   Both analog and digital methods have been used to transmit such information signals
over communication channels linking the source and user locations. Digital methods
tend to afford several advantages relative to analog techniques, including, for example,
improved immunity to channel noise and interference, increased capacity, and improved
security of communication through the use of encryption.

15   In transmitting an information signal from a source location over a
communication channel, the information signal is first converted into a form suitable
for efficient transmission over the channel. Conversion, or modulation, of the
information signal involves varying a parameter of a carrier wave on the basis of the
information signal in such a way that the spectrum of the resulting modulated carrier
20   is confined within the channel bandwidth. At the user location the original message
signal is replicated from a version of the modulated carrier received subsequent to
propagation over the channel. Such replication is generally achieved by using an
inverse of the modulation process employed by the source transmitter.

Modulation also facilitates multiplexing, i.e., the simultaneous transmission of
25   several signals over a common channel. Multiplexed communication systems will
generally include a plurality of remote subscriber units requiring intermittent service of
relatively short duration rather than continuous access to the communication channel.
Systems designed to enable communication over brief periods of time with a set of
subscriber units have been termed multiple access communication systems.

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A particular type of multiple access communication system is known as a spread spectrum system. In spread spectrum systems, the modulation technique utilized results in a spreading of the transmitted signal over a wide frequency band within the communication channel. One type of multiple access spread spectrum system is a code

5   division multiple access (CDMA) modulation system. Other multiple access communication system techniques, such as time division multiple access (TDMA), frequency division multiple access (FDMA) and AM modulation schemes such as amplitude companded single sideband are known in the art. However, the spread spectrum modulation technique of CDMA has significant advantages over these

10 modulation techniques for multiple access communication systems. The use of CDMA techniques in a multiple access communication system is disclosed in U.S. Pat. No. 4,901,307, issued Feb. 13, 1990, entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS", assigned to the assignee of the present invention.

15   In the above-referenced U.S. Pat. No. 4,901,307, a multiple access technique is disclosed where a large number of mobile telephone system users each having a transceiver communicate through satellite repeaters or terrestrial base stations using CDMA spread spectrum communication signals. In using CDMA communications, the frequency spectrum can be reused multiple times thus permitting an increase in system

20 user capacity. The use of CDMA results in a much higher spectral efficiency than can be achieved using other multiple access techniques.

More particularly, communication in a CDMA system between a pair of locations is achieved by spreading each transmitted signal over the channel bandwidth by using a unique user spreading code. Specific transmitted signals are extracted from

25 the communication channel by despreading the composite signal energy in the communication channel with the user spreading code associated with the transmitted signal to be extracted.

In particular spread spectrum communication systems it has been desired to allow various types of user channels (e.g., voice, facsimile, or high-speed data) to

30 operate at different data rates. These systems have typically been designed to have channels operative at a nominal data rate, and also to have reduced data rate traffic channels for providing more traffic data capacity. However, increasing traffic capacity

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by using reduced data rate channels lengthens the time required for data transmission. Moreover, in certain spread spectrum communication systems there is also a need for increased data rate traffic channels allowing for transmission at data rates higher than the nominal rate.

5   In order to support data transmission at variable rates it has generally been required to vary the rates of encoding, interleaving and modulation in accordance with the input data rate. This rate variation has typically required relatively sophisticated control of channel encoding and decoding processes, thereby increasing system cost and complexity

10   Accordingly, it is an object of the invention to provide a spread spectrum communication system in which communication channels are available for data transmission at both higher and lower than the nominal system rate.

          It is yet another object of the present invention to provide such a spread spectrum communication system in which a common format is used for encoding,
15   interleaving and modulating data to be transmitted at various rates.

          It is yet another object of the present invention to provide a CDMA spread spectrum communication system allowing for increases in traffic channel capacity in the absence of corresponding reductions in data rate.

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SUMMARY OF THE INVENTION

The implementation of CDMA techniques in spread spectrum communication systems using orthogonal PN code sequences reduces mutual interference between users, thereby allowing higher capacity and better performance. The present invention

5 provides an improved system and method for communicating information over in-phase (I) and quadrature phase (Q) communication channels in a CDMA spread spectrum communication system.

In an exemplary embodiment, an input information signal is transmitted over either an I or Q communication channel using a direct sequence spread spectrum

10 communication signal. The information signal is initially divided into first and second subsignals, which are respectively provided to first and second coset-encoding networks. The first coset-encoding combines the first subsignal with a first coset code, while the second coset-encoding network combines the second subsignal with a second coset code orthogonal to the first coset code. In this way the first and second coset-encoding

15 networks are operative to produce first and second coset-encoded signals, respectively. A composite coset-encoded signal formed from the first and second coset-encoded signals is then modulated by an orthogonal function signal to provide a first modulated signal.

In-phase pseudorandom noise (PN_I) and quadrature phase pseudorandom noise

20 (PN_Q) signals of predetermined PN codes are used for spreading the first modulated signal over either the I or Q communication channel, respectively. For example, the PN_I signal may be combined with the first modulated signal to provide an I-channel modulation signal for transmission to a receiver via the I communication channel.

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In the exemplary embodiment the receiver is operative to produce an estimate of the input information signal on the basis of the modulated carrier signal received over either the I or Q communication channel. The received signal is first demodulated using the orthogonal function signal. The demodulated signal is then decorrelated using
5   a despreading PN signal, with the resultant projection signals being provided to a phase rotator. The phase rotator operates to provide an estimate of the composite coset-encoded signal based on the projection signals and a received pilot signal. Estimates of the first and second subsignals are made by performing a further decorrelation based upon the orthogonality of the first and second coset codes.

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BRIEF DESCRIPTION OF THE DRAWINGS

Additional objects and features of the invention will be more readily apparent from the following detailed description and appended claims when taken in conjunction with the drawings, in which:

15   FIG. 1 shows a block diagram of a conventional spread spectrum transmitter.

FIG. 2 shows a block diagram of a preferred embodiment of a spread spectrum transmitter disposed to transmit I-channel and Q-channel information signals.

FIG. 3 shows a block diagrammatic representation of an I-channel coset encoding network operative to encode information signals in accordance with the
20   invention.

FIG. 4 is a block diagrammatic representation of a rate 1/p coset encoder of a type suitable for inclusion in the coset encoding network of FIG. 3.

FIG. 5 shows a block diagrammatic representation of a pair of I-channel and Q-channel coset encoding networks utilized in a preferred embodiment of the invention
25   to transmit data at four times a nominal data rate.

FIG. 6 shows a block diagrammatic representation of a pair of I-channel and Q-channel 1/4 rate coset encoding networks utilized in a preferred embodiment of the invention to transmit data at eight times the nominal rate.

FIG. 7 shows a block diagrammatic representation of a coset encoding network
30   utilized in a preferred embodiment to transmit data at a rate equivalent to one-half of the nominal rate.

FIG. 8 shows a block diagrammatic representation of a coset encoding network

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utilized in a preferred embodiment to transmit data at a rate equivalent to one-fourth of the nominal rate.

FIG. 9 depicts a pilot generation network for providing I and Q channel pilot sequences.

FIG. 10 shows an exemplary implementation of an RF transmitter incorporated within a preferred embodiment of the invention.

FIG. 12 is a block diagram of an exemplary diversity receiver disposed to receive the RF signal energy transmitted over the I and Q communication channels.

FIG. 13 is a block diagram of a receiver finger included within the diversity receiver of FIG. 12 designed to process signal energy received over a selected transmission path.

FIG. 14 provides a more detailed representation of the selected receiver finger illustrated in FIG. 13.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Referring to FIG. 1, there is shown a spread spectrum transmitter such as is described in U.S. Patent No. 5,103,459, issued 1992, entitled "SYSTEM AND METHOD FOR GENERATING SIGNAL WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM", which is assigned to the assignee of the present invention, and which is herein incorporated by reference. In the transmitter of FIG. 1, data bits 100 consisting of, for example, voice converted to data by a vocoder, are supplied to an encoder 102 where the bits are convolutional encoded with code symbol repetition in accordance with the input data rate. When the data bit rate is less than the bit processing rate of the encoder 102, code symbol repetition dictates that encoder 102 repeat the input data bits 100 in order to create a repetitive data stream at a bit rate which matches the operative rate of encoder 102. The encoded data is then provided to interleaver 104 where it is convolutional interleaved. The interleaved symbol data is output from interleaver 104 at an exemplary rate of 19.2 ksps to an input of exclusive-OR gate 106.

In the system of FIG. 1 the interleaved data symbols are scrambled to provide greater security in transmissions over the channel. Scrambling of the voice channel signals may be accomplished by pseudonoise (PN) coding the interleaved data with a

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PN code specific to an intended recipient subscriber unit. Such PN scrambling may be provided by the PN generator 108 using a suitable PN sequence or encryption scheme. The PN generator 108 will typically include a long PN generator for producing a unique PN code at a fixed PN chip rate of 1.2288 MHz. This PN code is then passed through

5 a decimator, with the resulting 19.2 kilo-symbol per second (ksps) scrambling sequence being supplied to the other input of exclusive-OR 106 in accordance with subscriber unit identification information provided thereto. The output of exclusive-OR 106 is then provided to one input of exclusive-OR 110.

Again referring to FIG. 1, the other input of exclusive-OR gate 110 is connected

10 to a Walsh waveform generator 112. Walsh generator 112 generates a Walsh waveform assigned to the data channel over which information is being transmitted. The Walsh waveform provided by generator 112 is selected from a set of 64 Walsh waveforms, each having a length of length 64 Walsh chips. The 64 orthogonal waveforms correspond to entries within a 64 by 64 Hadamard matrix wherein a particular Walsh

15 waveform is defined by a row or column of the matrix. The scrambled symbol data and Walsh waveform are exclusive-OR'ed by exclusive-OR gate 110 with the result provided as an input to both of the exclusive-OR gates 114 and 116.

Exclusive-OR gate 114 also receives a PN_I signal, while the other input of exclusive-OR gate 116 receives a PN_Q signal. The PN_I and PN_Q signals are

20 pseudorandom noise sequences typically corresponding to a particular area, i.e., cell, covered by the CDMA system and relate respectively to in-phase (I) and quadrature phase (Q) communication channels. The PN_I and PN_Q signals are respectively exclusive-OR'ed with the output of exclusive-OR gate 110 so as to further spread the user data prior to transmission. The resulting I-channel code spread sequence 122 and

25 Q-channel code spread sequence 126 are used to bi-phase modulate a quadrature pair of sinusoids. The modulated sinusoids are summed, bandpass filtered, shifted to an RF frequency, and again filtered and amplified prior to being radiated via an antenna to complete transmission over the communication channel.

Conventional techniques for accommodating variable data rates within the

30 transmission system of FIG. 1 have generally required utilization of a controller for varying the operating rates of the encoder 102, interleaver 104 and Walsh generator 112 in accordance with the input data rate. As is described hereinafter, the present

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invention enables spread spectrum transmission of an information signal at higher than
the nominal rate, or of transmission of a plurality of information signals at lower than
the nominal rate, using common encoding, interleaving and modulation rates.

FIG. 2 shows a block diagram of a preferred embodiment of a spread spectrum
5   transmitter 150 of the invention disposed to transmit an input information signal S_{IN} of
data rate kR_b, where k is an integer constant and R_b denotes a nominal transmitter data
(i.e., bit) rate. As employed herein the nominal data rate R_b is defined as being
equivalent to the product of the PN chip rate and the convolutional encoding code rate,
divided by the number of Walsh chips per symbol of the Walsh waveform. In an
10   exemplary embodiment a nominal transmitter data rate R_b of 9.6 kbps is set by utilizing
a set of modulation parameters in which the PN chip rate is selected to be 1.2288 MHz,
the convolutional code rate is a rate 1/2 code, and the Walsh waveform symbol length
is set at 64. It is a feature of the present invention that the transmitter 150 may be
employed to transmit information signals having data rates greater than or equal to the
15   nominal rate without adjustment of the values of the foregoing modulation parameters.
As is described hereinafter, the present invention also provides a technique for
transmitting a plurality of information signals of data rates lower than the nominal rate
without requiring corresponding modulation parameter adjustment.

In particular applications the input information bit sequence S_{IN} may consist of,
20   for example, voice converted to a stream of data bits by a vocoder. As is indicated by
FIG. 2, the input data stream is supplied to an encoding and interleaving network 160.
The network 160 convolutional encodes the information bit sequence S_{IN}, with the
encoded data then being interleaved and output from the network 160 as an encoded and
interleaved symbol stream S_{INT}. Assuming rate 1/2 convolutional encoding, the symbol
25   stream S_{INT} is supplied to a demultiplexer 170 at a symbol rate of $2kR_b$. The
demultiplexer 170 transforms the symbol stream S_{INT} into a set of k symbol substreams
{A(1), A(2), ... A(k)} each at a rate of $2R_b$, by routing successive symbols $S_{INT,i}$ to
successive ones of the substreams {A(1), A(2), ... A(k)}. The first k/2 symbol
substreams are provided to an I-channel coset-encoding network 180, while the
30   remaining k/2 symbol substreams are provided to a Q-channel coset-encoding network
190. As is described hereinafter, in exemplary implementations of the coset-encoding
networks 180 and 190 the symbol substreams are encoded using orthogonal sets of coset

codes of length p, where p = k/2. The coset-encoded symbol substreams within the networks 180 and 190 are then summed into I-channel and Q-channel composite symbol streams I_c and Q_c, respectively. Although for completeness both an I-channel and a Q-channel coset-encoding network are depicted in FIG. 2, in particular implementations

5 it may be desired to partition the symbol stream S_{INT} into only k/2 symbol substreams for transmission over either the I-channel or the Q-channel.

Referring again to FIG. 2, a pair of identical Walsh waveforms are provided to I-channel and Q-channel modulation and spreading networks 200 and 205 by a Walsh waveform generator 210. The Walsh waveforms are used within the networks 200 and

10 205 to modulate the I-channel and Q-channel composite symbol streams I_c and Q_c. In addition, PN spreading signals are also respectively provided to the modulation and spreading networks 200 and 205 by PN_I and PN_Q sequence generators 215 and 220. The PN_I sequence is used to spread the composite symbol stream I_c into an I-channel code spread sequence S_I. Similarly, the PN_Q sequence is utilized by the network 205

15 to spread the composite symbol stream Q_c into a Q-channel code spread sequence S_Q. The resultant I-channel and Q-channel code spread sequences S_I and S_Q are used to bi-phase modulate a quadrature pair of sinusoids generated within an RF transmitter 225. The modulated sinusoids will generally be summed, bandpass filtered, shifted to an RF frequency, and amplified prior to being radiated via an antenna over I and Q

20 communication channels.

FIG. 3 shows a block diagrammatic representation of the I-channel coset encoding network 180, it being understood that the Q-channel coset encoding network may be realized in a substantially identical manner. The encoding network 180 includes a plurality of coset encoders 250 to which are supplied the k/2 symbol substreams from

25 the demultiplexer 170. The encoders 250 are operative to generate k/2 sequences {a(1), a(2), ... a(k/2)} in which

$$a(1) = A(1)[\oplus]S_1$$

$$a(2) = A(2)[\oplus]S_2,$$

$$a(k/2) = A(k/2)[\oplus]S_{k/2}$$

30 where S_1, S_2, ... $S_{k/2}$ form a set of k/2 orthogonal coset codes of length p, and where the operation $[\oplus]$ is defined as follows. Let $A = (a_1, ..., a_r)$ be a sequence of length "r" and let $B = (b_1, b_2, ..., b_k)$, be a sequence of length "k", then $A [\oplus] B$ denotes the

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sequence $(a_1 \oplus b_1, \ldots, a_1 \oplus b_k, a_2 \oplus b_1 \ldots, a_2 \oplus b_k, \ldots, a_r \oplus b_k)$, where $\oplus$ denotes the exclusive-OR operation. In generating the sequences $\{a(1), a(2), \ldots a(k/2)\}$ each symbol within the symbol substreams $\{A(1), A(2), \ldots A(k)\}$ is repeated "p" times, with the "p^{th}" repeated symbol being exclusive-OR'ed with the pth coefficient of the

5   corresponding coset code. This operation has been characterized by those skilled in the art as encoding using a "rate $1/p$ repetition coset code".

FIG. 4 is a block diagrammatic representation of a rate $1/p$ coset encoder 300 disposed to use a coset code C to encode an input symbol stream R_s into an output coset-encoded symbol stream $R_{s,enc}$, where $C \in \{c_1, c_2, \ldots, c_p\}$. The coset encoder

10   includes a demulitplexer 305 for providing each symbol r, included within the symbol stream R_s to a set of p exclusive-OR gates 310. Each of the symbols r_i is exclusive-OR'ed with one of the coset code coefficients c_p, with the result being supplied to a p:1 multiplexer 315. The multiplexer 315 then produces the coset-encoded symbol stream $R_{s,enc}$, where $R_{s,enc} \in \{r_1 \oplus c_1, r_1 \oplus c_2, \ldots r_1 \oplus c_p, r_2 \oplus c_1,$

15   $r_2 \oplus c_2, \ldots r_2 \oplus c_p, \ldots, r_i \oplus c_p, \ldots\}$. More generally, for each symbol r_i the rate $1/p$ coset encoder produces a sequence,

$$(r_i \oplus c_1, r_i \oplus c_2, \ldots, r_i \oplus c_p) = r_i [\oplus] C.$$

Referring again to FIG. 3, in the preferred embodiment the substreams $\{A(1), A(2), \ldots A(k)\}$ and the coset codes $S_1, S_2, \ldots, S_{k/2}$ are composed of the logical values

20   0 and 1, as are the sequences $\{a(1), a(2), \ldots a(k/2)\}$ generated by the coset encoders 250. The sequences $\{a(1), a(2), \ldots a(k/2)\}$ are converted to an integer, i.e., ± 1, representation by a set of binary-to-integer conversion circuits 260 as follows:

$$0 \rightarrow +1$$

$$1 \rightarrow -1$$

25   As shown in FIG. 3, the sequence I_c is then created by combining the outputs from the conversion circuits 260 within a digital adder 270.

Embodiments Supporting High Data Rates

I. 4x Nominal Rate

30   FIG. 5 shows a block diagrammatic representation of a pair of I-channel and Q-channel coset encoding networks 350 and 360 utilized in a preferred embodiment of the invention to transmit data at four times the nominal rate. In particular, a rate $1/2$

encoded and interleaved symbol stream at a rate of eight times (e.g., 76.8 ksps) the nominal rate (e.g., 9.6 ksps) is demultiplexed by sequentially assigning symbols to one of four substreams $\{A(1), A(2), A(3), A(4)\}$, where $A(1) = \{A_{11}, A_{12}, ...\}$, $A(2) = \{A_{21}, A_{22}, ...\}$, $A(3) = \{A_{31}, A_{32}, ...\}$, and $A(4) = \{A_{41}, A_{42}, ...\}$. In the

5 implementation of FIG. 5 the rate 1/2 encoded and interleaved symbol stream is derived from an input data bit sequence (not shown) of a rate equivalent to four times the nominal rate. As is indicated by FIG. 5, the substreams $A(1)$ and $A(2)$ are respectively provided to rate 1/2 coset encoders 370 and 372 within the I-channel coset encoding network 350, while substreams $A(3)$ and $A(4)$ are respectively routed to rate

10 1/2 coset encoders 375 and 377 within the Q-channel coset encoding network 360. For rate 1/2 repetition the coset code (0,0) is used by the encoders 370 and 375 to encode the symbol substreams $A(1)$ and $A(3)$, while the coset code (0,1) is supplied to the coset encoders 372 and 377 for encoding the symbol substreams $A(2)$ and $A(4)$. The encoded substreams from the I-channel coset encoders 370 and 372 are transformed into an

15 integer format (± 1) by a pair of binary-to-integer conversion networks 380, and combined within digital adder 385 into the real sequence $I_{c,4}$. In like manner the substreams from the Q-channel coset encoders 375 and 377 are put into an integer format by the binary-to-integer conversion networks 390, and are then added within digital adder 395 to form the real sequence $Q_{c,4}$.

20 FIG. 5 also shows preferred implementations of the I-channel and Q-channel modulation and spreading networks 200 and 205. The I-channel network 200 includes a multiplier 400 for multiplying the sequences $I_{c,4}$ and $Q_{c,4}$ by a Walsh function W provided by the Walsh generator 210 in an integer (i.e., +/- 1 format), where in an exemplary implementation $W=(W_1,W_2,...,W_{32},W_{33},...,W_{64})$. In this way the coset

25 encoding networks 350 and 360 operate in conjunction with the spreading networks 200 and 205 to effectively assign the Walsh function W to the substreams $A(1)$ and $A(3)$, and to assign a Walsh function W^* to the substreams $A(2)$ and $A(4)$, where $W^*=(W_1,W_2,...,W_{32},-W_{33},...,-W_{64})$.

A PN_I sequence is provided to a multiplier 402 operative to spread the sequence

30 $I_{c,4}$ into the I-channel code spread sequence $S_{I,4}$ produced by the I-channel network 200. Similarly, a PN_Q sequence is used by multiplier 404 in spreading the sequence $Q_{c,4}$ into a Q-channel code spread sequence $S_{Q,4}$ produced by the network 205. The resultant

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I-channel and Q-channel code spread sequences $S_{I,4}$ and $S_{Q,4}$ are used to bi-phase modulate a quadrature pair of sinusoids generated within an RF transmitter (not shown).

5   II. 8x Nominal Rate

FIG. 6 shows a block diagrammatic representation of I-channel and Q-channel 1/4 rate coset encoding networks 450 and 460 utilized in a preferred embodiment of the invention to transmit data at eight times the nominal rate. An input bit sequence at

10 eight times the nominal rate is rate 1/2 encoded and interleaved into a symbol stream at sixteen times (e.g., 153.6 ksps) the nominal rate (e.g., 9.6 kbps) is demultiplexed by sequentially assigning symbols to one of eight substreams $A(i), i = 1,...,8$, where $A(i) = \{A_{i1}, A_{i2},....\}, i = 1....,8$.

As is indicated by FIG. 5, the substreams A(1)-A(4) are respectively provided

15 to I-channel rate 1/4 coset encoders 470, 472, 474 and 478 within the I-channel coset encoding network 450, while substreams A(5)-A(8) are respectively routed to Q-channel rate 1/4 coset encoders 480, 482, 484 and 488 within the Q-channel coset encoding network 460. A rate 1/4 coset code S_1 is used by the encoders 470 and 480 to encode the symbol substreams A(1) and A(5), a coset code S_2 is used by the encoders 472 and

20 482 to encode the symbol substreams A(2) and A(6), a coset code S_3 is used by the encoders 474 and 484 to encode the symbol substreams A(3) and A(6), while the coset code S_4 is used by the encoders 478 and 488 to encode the symbol substreams A(4) and A(8). The coset codes S_1 through S_4 are defined as follows:

$$S_1 = (s_{11}, s_{12}, s_{13}, s_{14}) = (0,0,0,0);$$

25 $$S_2 = (s_{21}, s_{22}, s_{23}, s_{24}) = (0,1,0,1);$$

$$S_3 = (s_{31}, s_{32}, s_{33}, s_{34}) = (0,0,1,1,): \text{ and}$$

$$S_4 = (s_{41}, s_{42}, s_{43}, s_{44}) = (0,1,1,0).$$

In this way the eight coset encoders operate to produce a set of eight encoded symbol streams a(i), where i = 1, ..., 8, at a rate (e.g. 76.8 ksps) equivalent to eight times the

30 nominal rate. The encoded symbol streams a(i) are produced in accordance with the following expression:

$$a(i) = \begin{cases} A(i)[\oplus]S_i & 1 \le i \le 4 \\ A(i)[\oplus]S_{i-4} & 5 \le i \le 8 \end{cases}.$$

In order to simplify notation without loss of generality, it will be assumed hererinafter that each substream $A(i)$ is composed of a single symbol A_i rather than the sequence A_{ij}, where the subscript "j" represents time. For example, using this notation to define $a(8)$ yields,

5   $a(8) = A_x[\oplus]S_x = \{A_x \oplus 0, A_x \oplus 1, A_x \oplus 1, A_x \oplus 0\}$

The sequences $a(i)$, $i = 1, ..., 8$, are then converted by binary-to-integer converters 490 to a set of real sequences $r(i)$, $i = 1, ..., 8$, given by

$$r(i) = (-1)^{a(i)} = ((-1)^{a_{i1}}, ..., (-1)^{a_{ip}}) = (r_{i1}, ..., r_{ip}).$$

where $a_{ij} = A_i s_{ij}$, and where s_{ij} denotes the jth symbol included within the ith coset code

10   S_i. The sequences $r(i)$, $i = 1, ..., 4$, are combined within digital adder 494 into the real sequence $I_{c,8}$. In like manner the real sequences $r(i)$, $i = 5, ..., 8$, are added within digital adder 498 to form the real sequence $Q_{c,x}$. Referring to FIG. 6, multipliers 502 and 504 are provided for multiplying the sequences $I_{c,x}$ and $Q_{c,x}$ by a Walsh function W provided by the Walsh generator 506, where in an exemplary implementation

15   $W = (W_1, W_2, ..., W_{32}, W_{33}, ..., W_{64})$. In this way the Walsh functions $W0$, W^1, W^2, W^3 are effectively assigned to the symbol substreams $A(i)$, $i = 1, ..., 4$, and to $A(i)$, $i = 5$, ..., 8, respectively, where W^0, W^1, W^2, W^3 are defined as:

$W^0 = (W_a, W_b, W_c, W_d)$;

$W^1 = (W_a, -W_b, W_c, -W_d)$;

20   $W^2 = (W_a, W_b, -W_c, -W_d)$; and

$W^3 = (W_a, -W_b, -W_c, W_d)$.

The sequences W_a, W_b, W_c, W_d may be defined in terms of the Walsh waveform W as:

$W_a = (W_1, ..., W_{16})$;

$W_b = (W_{17}, ..., W_{32})$;

25   $W_c = (W_{33}, ..., W_{48})$; and

$W_d = (W_{49}, ..., W_{64})$.

A PN_1 sequence is provided to a multiplier 510 operative to spread the sequence

$I_{c,s}$ into an I-channel code spread sequence $S_{I,s}$. Similarly, a PN_Q sequence is used by multiplier 514 in spreading the real sequence $Q_{c,s}$ into a Q-channel code spread sequence $S_{Q,s}$. The resultant I-channel and Q-channel code spread sequences $S_{I,s}$ and $S_{Q,s}$ are used to bi-phase modulate a quadrature pair of sinusoids generated within an

5 RF transmitter (not shown).

Embodiments Supporting Lower Data Rates

1. 1/2 Data Rate

Referring to FIG. 7. a pair of input data streams $A_{nom/2}$ and $B_{nom/2}$ are supplied

10 at a data rate equivalent to one-half of the nominal rate to encoding and interleaving networks 550 and 554. The networks 550 and 554 convolutional encode the signals $A_{nom/2}$ and $B_{nom/2}$ into encoded and interleaved symbol streams $A_{1/2}(1)$ and $A_{1/2}(2)$, where $A_{1/2}(1) = \{A_{11}, A_{12}, ...\}$, and $A_{1/2}(2) = \{A_{21}, A_{22}, ...\}$. Assuming rate 1/2 convolutional encoding, the resultant interleaved symbol streams $A_{1/2}(1)$ and $A_{1/2}(2)$ are

15 supplied at the nominal rate to coset encoders 558 and 560. The coset code S_1, where $S_1 = (0,0)$, is used by the encoder 558 to rate 1/2 repetition encode the symbol substream $A_{1/2}(1)$ into an encoded substream $a_{1/2}(1)$. In similar manner the coset code S_2, where $S_2 = (0,1)$, is supplied to the coset encoder 560 for rate 1/2 repetition encoding of the symbol substream $A_{1/2}(2)$ into an encoded substream $a_{1/2}(2)$. The

20 encoded substreams $a_{1/2}(1)$ and $a_{1/2}(2)$ are defined as:

$$a_{1/2}(1) = A_{1/2}(1) \; [\oplus] \; S_1 = \{A_{11} \oplus 0, A_{11} \oplus 0, ...\}, \text{ and}$$

$$a_{1/2}(2) = A_{1/2}(2) \; [\oplus] \; S_2 = \{A_{21} \oplus 0, A_{21} \oplus 1, ...\}.$$

The encoded substreams are output from the coset encoders 558 and 560 at twice the nominal rate and are transformed into an integer format (± 1) by a pair of

25 binary-to-integer conversion networks 570. The resulting real sequences $r_j(1)$ and $r_j(2)$ are combined within digital adder 575 into the real sequence $R_{1/2}$ for subsequent transmission to a jth receiver area. The real sequence $R_{1/2}$ is provided to a multiplier 580 for multiplication by a Walsh function W provided by a Walsh generator 590, where in an exemplary implementation $W=(W_1, W_2, ..., W_{32}, W_{33}, ..., W_{64})$. This results

30 in assignment of the Walsh function (W, W) to the symbol stream $A_{1/2}(1)$, and assignment of the Walsh function W^* to symbol stream $A_{1/2}(2)$, where $W^* = (W, -W)$. Following multiplication by the Walsh function W, the sequence $R_{1/2}$ will generally be

spread by a pseudorandom PN_I or PN_Q sequence for RF transmission over either a corresponding in-phase (I) or quadrature phase (Q) communication channel.

II. 1/4 Data Rate

Referring to FIG. 8, a set four input data streams $A_{nom/4}$, $B_{nom/4}$, $C_{nom/4}$ and $D_{nom/4}$ are supplied at a data rate equivalent to one-fourth of the nominal rate to encoding and interleaving networks 601, 602, 603 and 604. The networks 601-604 convolutional encode the data streams $A_{nom/4}$, $B_{nom/4}$, $C_{nom/4}$ and $D_{nom/4}$ into encoded and interleaved symbol streams $A_{1/4}(1)$, $A_{1/4}(2)$, $A_{1/4}(3)$ and $A_{1/4}(4)$, where

$$A_{1/4}(1) = \{A_{11}, A_{12}, \ldots\},$$
$$A_{1/4}(2) = \{A_{21}, A_{22}, \ldots\},$$
$$A_{1/4}(3) = \{A_{31}, A_{32}, \ldots\}, \text{ and}$$
$$A_{1/4}(4) = \{A_{41}, A_{42}, \ldots\}.$$

Assuming rate 1/2 convolutional encoding, the resultant interleaved symbol streams $A_{1/4}(1)$, $A_{1/4}(2)$, $A_{1/4}(3)$ and $A_{1/4}(4)$ are supplied at one-half of the nominal rate to coset encoders 611, 612, 613 and 614. The coset codes $\{(0000), (0101), (0011), (0110)\}$ are respectively used by the coset encoders 611-614 to encode the symbol streams $A_{1/4}(1)$, $A_{1/4}(2)$, $A_{1/4}(3)$ and $A_{1/4}(4)$ into the encoded substreams $a_{1/4}(1)$, $a_{1/4}(2)$, $a_{1/4}(3)$, and $a_{1/4}(4)$. The substreams $a_{1/4}(1)$, $a_{1/4}(2)$, $a_{1/4}(3)$, and $a_{1/4}(4)$ may be represented as:

$$a_{1/4}(1) = \{A_{11}\oplus 0, A_{11}\oplus 0, A_{11}\oplus 0, A_{11}\oplus 0, \ldots\},$$
$$a_{1/4}(2) = \{A_{21}\oplus 0, A_{21}\oplus 1, A_{21}\oplus 0, A_{21}\oplus 1, \ldots\},$$
$$a_{1/4}(3) = \{A_{31}\oplus 0, A_{31}\oplus 0, A_{31}\oplus 1, A_{31}\oplus 1, \ldots\}, \text{ and}$$
$$a_{1/4}(4) = \{A_{41}\oplus 0, A_{41}\oplus 1, A_{41}\oplus 1, A_{41}\oplus 0, \ldots\}.$$

The encoded substreams are output from the coset encoders 611-614 at twice the nominal rate and are transformed into an integer format (± 1) by binary-to-integer conversion networks 620. A set of resulting real $r_j(i)$, $i = 1$ to 4, sequences for transmission to a jth receiver are combined within digital adder 575 into the real sequence $R_{1/4}$. The real sequence $R_{1/4}$ is provided to a multiplier 624 for multiplication by a Walsh function W_j associated with the jth receiver. The sequence W_j is provided by a Walsh generator 630, and is defined as $W_j=(W_{j1},W_{j2},\ldots,W_{j32},W_{j33},\ldots,W_{j64})$. This results in assignment of the Walsh functions W^0, W^1, W^2, W^3 to the symbol streams $A_{1/4}(1)$, $A_{1/4}(2)$, $A_{1/4}(3)$ and $A_{1/4}(4)$, where W^0, W^1, W^3, W^4 are given by:

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$$W^0 = (W_j, \ W_j, \ W_j, \ W_j);$$

$$W^1 = (W_j, \ -W_j, \ W_j, \ -W_j);$$

$$W^2 = (W_j, \ W_j, \ -W_j, \ -W_j); \ \text{and}$$

$$W^3 = (W_j, \ -W_j, \ -W_j, \ W_j).$$

5 It is thus apparent that four distinct information signals. respectively identified by the Walsh waveforms W^0, W^1, W^3, W^4, are capable of being transmitted to the jth receiver through utilization of a single Walsh waveform W_j in conjunction with the coset-encoding technique contemplated by the invention. Subsequent to being multiplied by the Walsh function W_j. the sequence $R_{1,j}$ will typically be spread by a pseudorandom

10 PN_I or PN_Q sequence for RF transmission over either a corresponding in-phase (I) or quadrature phase (Q) communication channel.

Assuming transmission over the I channel to the jth user, the transmitted sequence synthesized from the sequences $r_j(i)$ may be represented as

$$S_{Ij} = \left(\sum_{i=1}^{p} r_j(i) \right) \cdot W_j \cdot PN_I.$$

15 where p=4 in the example of FIG. 8. If transmission were instead to occur over the Q channel, the transmitted sequence would be represented as:

$$S_{Qj} = \left(\sum_{i=p+1}^{k} r_j(i) \right) \cdot W_j \cdot PN_Q.$$

Exemplary sets of parameters used in supporting transmission of input symbol streams at various data rates (R_b) are summarized below in Table I. For each data rate

20 Table I provides a corresponding input symbol repetition rate. repetition coset code rate, as well as Walsh waveform length and chip rate. Each entry (X-Y) within the "Demux" column specifies the number of input symbol streams (X) at the associated data rate R_b, and the number of symbol substreams (Y) into which the input symbol stream(s) are demultiplexed for coset-encoding.

25

Table I

R_b [kbs]	Rate 1/p Input Symbol Repetition	I/Q Channel	Demux	Rate 1/p Coset Code Repetition	Walsh Length	Walsh Code Rate [Mcps]
9.6	1	I or Q	1-1	1	64	1.2288
4.8	1/2	I or Q	2-2	1/2	64	1.2288
2.4	1/4	I or Q	4-4	1/4	64	1.2288
1.2	1/8	I or Q	8-8	1/8	64	1.2288
19.2	1	I and Q	1-2	1	64	1.2288
38.4	1	I and Q	1-4	1/2	64	1.2288
76.8	1	I and Q	1-8	1/4	64	1.2288
19.2	1	I or Q	1-1	1	128	2.457
9.6	1/2	I or Q	2-2	1/2	128	2.457
4.8	1/4	I or Q	4-4	1/4	128	2.457
2.4	1/8	I or Q	8-8	1/8	128	2.457
1.2	1/16	I or Q	16-16	1/16	128	2.457
38.4	1	I and Q	1-2	1	128	2.457
76.8	1	I and Q	1-4	1/2	128	2.457
153.6	1	I and Q	1-8	1/4	128	2.457
38.4	1	I or Q	1-1	1	256	4.9
19.2	1/2	I or Q	2-2	1/2	256	4.9
9.6	1/4	I or Q	4-4	1/4	256	4.9
4.8	1/8	I or Q	8-8	1/8	256	4.9
2.4	1/16	I or Q	16-16	1/16	256	4.9
1.2	1/32	I or Q	32-32	1/32	256	4.9
38.4	1	I and Q	1-2	1	256	4.9
76.8	1	I and Q	1-4	1/2	256	4.9
153.6	1	I and Q	1-8	1/4	256	4.9

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Transmission of the Coset-Encoded Data Over the I and Q channels

In the preferred embodiment, a pilot channel containing no data modulation is transmitted together the "N" receivers within a given cell or sector together with the I-channel and Q-channel spread sequences S_{Ij} and S_{Qj}, $j = 1$ to N The pilot channel may

5   be characterized as an unmodulated spread spectrum signal used for signal acquisition and tracking purposes.   In systems incorporating a plurality of transmitters in accordance with the invention, the set of communication channels provided be each will be identified by a unique pilot signal. However, rather than using a separate set of PN generators for the pilot signals, it is realized that a more efficient approach to

10   generating a set of pilot signals is to use shifts in the same basic sequence. Utilizing this technique an intended receiver unit sequentially searches the whole pilot sequence and tunes to the offset or shift that produces the strongest correlation.

Accordingly, the pilot sequence will preferably be long enough that many different sequences can be generated by shifts in the basic sequence to support a large

15   number of pilot signals in the system. In addition, the separation or shifts must be great enough to ensure that there is no interference in the pilot signals. Hence, in an exemplary embodiment the pilot sequence length is chosen to be 2^{15}, which allows for 512 distinct pilot signals with offsets in a basic sequence of 64 chips.

Referring to FIG. 9, a pilot generation network 630 includes a Walsh generator

20   640 for providing the Walsh "zero" W_o waveform consisting of all zeroes to digital multipliers 644 and 646. The Walsh waveform W_o is multiplied by the PN_I and PN_Q sequences provided by PN_I and PN_Q generators 647 and 648 to the multipliers 644 and 646, respectively. Since the waveform W_o includes only ones, the information content of the resultant sequences depends only upon the PN_I and PN_Q sequences. The

25   sequences produced by multipliers 644 and 646 are provided as inputs to Finite Impulse Response Filters (FIR) filters 650 and 652. The filtered sequences output from FIR filters 650 and 652, respectively corresponding to I-channel and Q-channel pilot sequences P_{Io} and P_{Qo}, are supplied to the RF transmitter 660 (FIG. 10).

Referring to FIG. 10, there is shown an exemplary implementation of the RF

30   transmitter 660. Transmitter 660 includes an I-channel summer 670 for summing the set of PN_I spread data signals S_{Ij}, $j = 1$ to N, with the I-channel pilot P_{Io} for transmission to the N receivers within a specified cell or sector. Similarly, a Q-channel summer 672

serves to combine the set of PN_Q spread data signals S_{Qj}, $j = 1$ to N, with the Q-channel pilot P_{Qo}. Digital to analog (D/A) converters 674 and 676 are provided for converting the digital information from the I-channel and Q-channel summers 670 and 672, respectively, into analog form. The analog waveforms produced by D/A converters 674 and 676 are provided along with local oscillator (LO) carrier frequency signals $Cos(2\pi ft)$ and $Sin(2\pi ft)$, respectively, to mixers 688 and 690 where they are mixed and provided to summer 692. The quadrature phase carrier signals $Sin(2\pi ft)$ and $Cos(2\pi ft)$ are provided from suitable frequency sources (not shown). These mixed IF signals are summed in summer 692 and provided to mixer 694.

Mixer 694 mixes the summed signal with an RF frequency signal from frequency synthesizer 696 so as to provide frequency upconversion to the RF frequency band. The RF signal includes in-phase (I) and quadrature phase (Q) components, and is bandpass filtered by bandpass filter 698 and output to RF amplifier 699. Amplifier 699 amplifies the band limited signal in accordance with an input gain control signal from transmit power control circuitry (not shown). It should be understood that differing implementations of the RF transmitter 630 may employ a variety of signal summing, mixing, filtering and amplification techniques not described herein, but which are well known to those in the art.

FIG. 11 is a block diagram of an exemplary diversity receiver disposed to receive the RF signal provided by the RF transmitter 630. In FIG. 11 the transmitted RF signal is received by antenna 710 and provided to a diversity RAKE receiver which is comprised of analog receiver 712 and digital receiver 714. The signal as received by antenna 710 and provided to analog receiver 712 may be comprised of multipath propagations of the same pilot and data signals intended for individual or multiple subscriber receivers. Analog receiver 712. which is configured in the exemplary embodiment as a QPSK modem, frequency downconverts, and digitizes the received signal into composite I and Q components. The composite I and Q components are provided to digital receiver 714 for demodulation. The demodulated data is then provided to digital circuitry 716 for combining, deinterleaving and decoding.

Each I and Q component output from analog receiver 712 may be comprised of multipath propagations of an identical pilot and corresponding information signals. In digital receiver 714 certain multipath propagations of the transmitted signal, as selected

by a searcher receiver 715 in combination with a controller 718, are each processed by a different one of multiple data receivers or demodulators 720a-720c, which are also referred to as "fingers". Although only three data demodulating fingers (demodulators 720a-720c) are illustrated in FIG. 11, it should be understood that more or less fingers

5 may be used. From the composite I and Q components each finger extracts, by despreading. the I and Q components RI and RQ of the pilot and data signals corresponding to a particular path.

The I and Q components of the pilot signal for each finger may be said to form a pilot vector. and the I and Q components of the I-channel and Q-channel data to form

10 a pair of data vectors. In accordance with the invention, these I and Q components of the pilot and data vectors are extracted from the received signal energy in order to produce estimates of the I-channel and Q-channel data. The pilot signal is typically transmitted at a greater signal strength than the data signals, and as such the magnitude of the pilot signal vector is greater than the received data signal vectors. Accordingly,

15 the pilot signal vector can be used as an accurate phase reference for signal processing.

In the transmission process the pilot and data signals as transmitted travel the same path to the receiver. However, due to channel noise the received signal will generally be offset from the transmitted phase angle. The formulation of the dot, i.e.. scalar, products of the pilot signal vector with I-channel and Q-channel data signal

20 vectors are used as disclosed herein to extract the I-channel and Q-channel data from the signal received by the selected receiver finger. In particular, the dot product is used to find the magnitudes of the components of the data vectors that are in phase with the pilot vector by projecting the pilot vectors onto each of the data vectors. One procedure for extracting the pilot signal from the signal energy received by the selected

25 receiver finger is described below with reference to FIG. 8, and also in copending United States Patent Application Serial No. 07/981,034, filed November 24, 1992, entitled "PILOT CARRIER DOT PRODUCT CIRCUIT", which is assigned to the assignee of the present invention, and which is herein incorporated by reference.

30 Recovery of Coset-Encoded Symbol Substreams

In what follows there will be described in detail the recovery from the I-channel transmitted data a single coset-encoded substream, a(i), where

21

$$a(i) = A(i) \oplus S_i = (A_{i1} \oplus s_{i1}, ..., A_{i1} \oplus s_{ip}).$$

It is assumed that prior to transmission to a jth (FIG. 8) of "N" receivers over the I and Q channels the substream a(i) is converted to a real sequence r(i), where

$$r(i) = (-1)^{a(i)} = ((-1)^{A_{i1} \oplus s_{i1}}, ..., (-1)^{A_{i1} \oplus s_{ip}}) = (r_{i1}, ..., r_{ip}).$$

5   After spreading by the Walsh waveform W_j and by the sequences PN_I and PN_Q, respectively, the resultant sequences S_{Ij} and S_{Qj} intended for reception by the jth receiver may be represented as

$$S_{Ij} = \sum_{i=1}^{p} r(i) \, [W_j \cdot PN_I],$$

and

$$S_{Qj} = \sum_{i=p+1}^{k} r(i) \, [W_j \cdot PN_Q],$$

10

The composite signal transmitted to the "N" receivers within a particular cell is given by

$$S(t) = \tilde{I}cos(\omega_o t) - \tilde{Q}sin(\omega_o t);$$

where

$$\tilde{I} = \sum_{j=1}^{N} s_{Ij}; \quad \tilde{Q} = \sum_{j=1}^{N} S_{Qj}$$

15   For clarity of presentation it will be assumed that the signal S(t) propagates over an mth transmission path to the jth receiver , which allows the signal $R_j(t)$ received thereby to be expressed as:

$$R_j(t) = \bar{I}\cos(\omega_o t + \theta) - \bar{Q}\sin(\omega_o t + \theta) + n(t);$$

where the signal $R_j(t)$ has a random phase shift of θ relative to the local reference of the receiver, and where n(t) denotes the inherent signal interference noise.

Referring to the block diagrammatic representation of FIG. 12, the jth receiver
5   is seen to include a set of "r" demodulating fingers 720 disposed to process the signal $R_j(t)$ as received over "r" transmission paths. The signal $R_j(t)$ transmitted over the mth path is passed through a bandpass filter having a transfer function h(t), and is sampled at times $t = kT_w$, where T_w denotes the period between successive chips in the assigned Walsh waveform W_j. These operations produce the I and Q projections $R_{Im,k}$ and $R_{Qm,k}$
10 supplied to the mth demodulating finger 720, where

$$R_{Im,k} = R_j(t)\cos(\omega_o t)^*h(t)\Big|_{t=kT_w + \tau_m} = \bar{I}\cos\theta - \bar{Q}\sin\theta + N_i;$$

$$R_{Qm,k} = -R_j(t)\sin(\omega_o t)^*h(t)\Big|_{t=kT_w + \tau_m} = \bar{I}\sin\theta + \bar{Q}\cos\theta + N_q$$

where τ_m corresponds to the delay associated with the mth transmission path, and where the noise terms N_i and N_q may be characterized as random processes of zero mean and
15 variance σ^2. In accordance with the invention, estimates of the sequence r(i) transmitted over the mth transmission path are derived from the sampled projections $R_{Im,k}$ and $R_{Qm,k}$ by the mth receiver finger 720.

Referring to FIG. 13, there is shown a block diagram of the mth receiver finger 720 operative to process the sampled projections $R_{Im,k}$ and $R_{Qm,k}$. The receiver finger
20 720 includes a demodulation/despreading and phase rotation circuit 740, as well as a phase estimation and time tracking circuit 744. In accordance with the invention, the circuit 740 operates to demodulate the sampled projections $R_{Im,k}$ and $R_{Qm,k}$ by performing a first set of partial correlations using the assigned Walsh waveform W_j and PN_I sequence, and a second set of partial correlations using the assigned Walsh
25 waveform and the PN_Q sequence. Each partial correlation is performed over an interval of L/p Walsh chips, where L denotes the length of the Walsh waveform W_j used to

cover the "p" symbol substreams inherent within the sequences S_{Ij} and S_{Qj}. The results of the partial correlations are then rotated in phase in order to produce the decision variables Ihat(m) and Qhat(m) output by the mth receiver finger 720. This phase rotation is performed in accordance with an estimated phase shift $\hat{\theta}$ between the

5   transmitted waveform and a locally-generated reference. In a preferred implementation the phase estimation and time tracking circuit 744 includes a phase-locked for generating the phase estimate $\hat{\theta}$.

The phase estimation and time tracking circuit 744 operates to provide an estimate of the pilot signal (P_m) transmitted over the mth path on the basis of

10   intermediate signals produced by the circuit 740 during demodulation and despreading of the sampled projections $R_{Im,k}$ and $R_{Qm,k}$. The extracted pilot signal is used for phase rotation of the partial correlations within circuit 740, as well as for time alignment within a sample combiner 750 (FIG. 12). The results of these independent correlations are used to produce the mth pair of decision variables I(hat)(m) and Q(hat)(m) provided

15   to a sample combiner 750 (FIG. 12). Within the sample combiner 750 the decision variables Ihat(l), l=1 to r, produced by the set of "r" receiver fingers 720 are time-aligned and combined, as are the decision variables Qhat(m)

Referring to FIG. 14, the mth receiver finger 720 is seen to include multipliers 780 and 782 for receiving the sampled projections $R_{Im,k}$ and $R_{Qm,k}$ at the PN spreading

20   rate of 1.2288 MHz. A Walsh generator 786 is connected to both of multipliers 780 and 782, where its output (W_j) is multiplied with the projections $R_{Im,k}$ and $R_{Qm,k}$. The receiver finger 720 further includes PN generators 790 and 792 for providing the PN_I sequence to multipliers 798 and 800, and the PN_Q sequence to multipliers 802 and 804. As is indicated by FIG. 14, the Walsh demodulated projections $R'_{Im,k}$ and $R'_{Qm,k}$ from

25   multiplier 780 are multiplied with the PN_I sequence at multiplier 798 and with the PN_Q sequence at multiplier 802. Similarly, the output from multiplier 782 is multiplied with the PN_I sequence at multiplier 800, and with the PN_Q sequence at multiplier 804.

The multipliers 798 and 800 correlate the Walsh demodulated projections $R'_{Im,k}$ and $R'_{Qm,k}$ with the PN_I sequence. Appropriate timing is maintained between the PN_I

30   sequence and the sequences $R'_{Im,k}$ and $R'_{Qm,k}$ by a time alignment circuit 810, the operation of which is discussed below. Similarly, the sequences $R'_{Im,k}$ and $R'_{Qm,k}$ are correlated with the PN_Q sequence by multipliers 802 and 804. The correlated outputs

24

of multipliers 798, 800, 802 and 804 are then provided to corresponding I-channel accumulators 814 and 816, and Q-channel accumulators 818 and 820. Accumulators 814, 816, 818 and 820 accumulate the input information over L/p Walsh chips, where, again, L denotes the length of the Walsh waveform W_j. The accumulators 814, 816,

5 818 and 820 are operative to produce partial correlations A_{In}, A_{Qn}, B_{In}, and B_{Qn} during each of the "p" partial correlation intervals of length L/p Walsh chips (i.e., n = 1 to p) occurring during each Walsh waveform. The partial correlations A_{In}, A_{Qn}, B_{In}, and B_{Qn} are provided to delay elements 824, 826, 828 and 830 through corresponding switches 834, 836, 838 and 840. The switches are closed from normally-open positions at the

10 conclusion of each partial correlation interval in accordance with timing signals provided by the time alignment circuit 810. The partial correlations A_{In} and A_{Qn} produced by the I-channel accumulators 814 and 816 at the conclusion of the nth correlation interval may be expressed as:

$$A_{In} = \sum_{k=\frac{64}{p}n+1}^{\frac{64}{p}(n+1)} R_{Ik}W_{jk}PN_{Ik} = \frac{64}{p} \sum_{j=1}^{P} r_{jn}\cos\theta,$$

$$A_{Qn} = \sum_{k=\frac{64}{p}n+1}^{\frac{64}{p}(n+1)} R_{Qk}W_{jk}PN_{IK} = \frac{64}{p} \sum_{j=1}^{p} r_{jn}\sin\theta, \quad n=1, ..., p,$$

15

it being understood that the partial correlations B_{In} and B_{Qn} may be represented in substantially similar manner. Referring to equations (12) and (13), the terms r_{jn}, j = 1 to p, collectively represent an estimate of the "p" integral values included within the real sequence r(i) defined by equation (5).   Referring again to FIG. 14, the phase

20 estimation and time tracking circuit 744 includes a pilot extraction circuit 850 for producing pilot phase signals used in maintaining time alignment within the receiver finger 720. The pilot extraction circuit 850 includes a multiplier 854 to which is provided the outputs from multipliers 798 and 802, as well as a multiplier 856 for

multiplying the outputs of multipliers 800 and 804. The circuit 850 further includes Walsh generators 862 and 864 operative to supply the Walsh waveforms W_i and W_o, respectively, to a multiplier 866. The resultant demodulating waveform W_iW_o produced by multiplier 866, appropriately time-aligned by virtue of the timing information provided by circuit 810 to Walsh generators 862 and 864, is provided to multipliers 868 and 870. The waveform W_iW_o is multiplied with the output of multiplier 854 by multiplier 868, while multiplier 870 performs the same operation in response to the waveform W_iW_o and the output provided by multiplier 856.

The outputs of multipliers 868 and 870 are respectively accumulated by pilot extraction accumulators 874 and 878 over an interval selected to ensure generation of an unbiased estimate of the phase of the received pilot signal. In an exemplary embodiment the accumulation interval spans a time period of duration 2rL, where as noted above L corresponds to the Walsh symbol period. This accumulation interval will generally take place over the time periods of length "rL" occurring immediately before and after the time at which it is desired to estimate the pilot phase. Time alignment between the outputs produced by accumulators 814, 816, 818 and 820 and the outputs of pilot extraction accumulators 874 and 880 is maintained by the delay elements 824, 826, 828 and 830. The signal delay effected by each of the delay elements 824, 826, 828 and 830 is chosen to be of a duration equivalent to the interval spanned by the "r" future Walsh symbols. Accordingly, in generating the pilot estimate corresponding to the nth partial correlations A_{In} and A_{Qn} a set of data samples D_j, where $(L/p)(n-r) + 1 \leq j \leq (L/p)(n+r)$, are accumulated by the accumulators 874 and 878.

The signals produced by the pilot extraction accumulators 882 and 886 correspond to I-channel and Q-channel projections of the pilot (P_m) signal transmitted over the mth path, and may be respectively represented as:

$$P_m \cdot \cos(\hat{\theta}) = \sum_{(L/p)(n-r)+1}^{(L/p)(n+r)} \left\{ R_{Im}PN_IW_0 + R_{Qm}PN_QW_0 \right\} \qquad (15)$$

$$P_m \cdot \sin(\hat{\theta}) = \sum_{(L/p)(n-r)+1}^{(L/p)(n+r)} \left\{ -R_{Im}PN_QW_0 + R_{Qm}PN_IW_0 \right\} \qquad (16)$$

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Referring to FIG. 14, the I-channel and Q-channel projections of the pilot signal are each provided to both the I-channel phase rotator 850 and the Q-channel phase rotator 852. The I-channel phase rotator 850 produces a sequence of output data values corresponding to an estimate of the sequence r(t) transmitted over the mth path weighted

5 by the pilot signal P_m. The decision term $\hat{I}_n(m)$ generated by the I-channel phase rotator 850 at the conclusion of the nth correlation interval may be represented as:

$$\hat{I}_n(m) = A_{In} \cdot P_m \cdot \cos(\hat{\theta}) + A_{Qn} \ P_m \ \sin(\hat{\theta}) = \frac{64}{p} \sum_{i=1}^{p} r_{in}.$$

The sample combiner 750 (FIG. 12) combines the I-channel decision terms $\hat{I}_n(i)$, i=1 to r, produced by the finger demodulators 720 during the nth correlation

10 interval into a composite decision term $\hat{I}_{cn}$, and combines the Q-channel decision variables $\hat{Q}_n(i)$, into a composite decision term $\hat{Q}_{cn}$. The composite decision terms $\hat{I}_{cn}$ and $\hat{Q}_{cn}$ are serially output by the combiner 750 as the sequences,

$\hat{I}_c = (\hat{I}_1, ..., \hat{I}_p)$, and

$\hat{Q}_c = (\hat{Q}_1, ..., \hat{Q}_p)$,

15 where the subscripts indicate correspondence to the "p" symbol substreams combined into the real sequence r(i). The composite decision sequences $\hat{I}_c$ and $\hat{Q}_c$ are provided to I-channel and Q-channel multiplexers 870 and 874, which respectively produce the parallel outputs

$\hat{I}_c^T = (\hat{I}_1, ..., \hat{I}_p)^T$, and

20      $\hat{Q}_c^T = (\hat{Q}_1, ..., \hat{Q}_p)^T$.

In accordance with the invention, a set of estimates $\hat{A}_I(i)$ of the input symbol streams $A_I(i)$ transmitted over the I-channel, where i=1 to p, is generated by performing a decorrelation of the decision sequence $\hat{I}_c$ based on the coset codes inherent within the sequence r(i). More specifically, an estimate of the ith symbol stream $A_I(i)$ is made

25 through the following computation of the inner product of r(i) with the decision sequence $\hat{I}_I(i)$:

$$\hat{A}(i) = \sum_{n=1}^{p} \hat{I}_n(i)(-1)^{c_{i,n}} = 64(-1)^{A(i)}$$

where $c_{i,n}$ denotes the nth term of the coset code c_i used to encode the ith symbol stream. The computation specified by equation (18) relies on the orthogonality

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between the coset codes used to encode the input symbol streams. That is,

$$\hat{A}(i) = \sum_{n=1}^{P} \hat{I}_n(i)(-1)^{q_n} = 0$$

for all $j \neq j$. For $p \geq 4$ equation (18) may be solved by performing, for example, a Fast Hadamard Transform (FHT) upon the upon the sequences $\hat{I}_c^T$ provided the
5   multiplexer 870 (FIG. 12). The symbol stream estimates are then deinterleaved and decoded in order to estimate the transmitted data.

The previous description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. The various modifications to these embodiments will be readily apparent to those skilled in the art,
10 and the generic principles defined herein may be applied to other embodiments without the use of inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

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WHAT IS CLAIMED IS:

1. A transmitter for modulating an information signal for transmission in a
2 spread spectrum communication system comprising:

means for demultiplexing said information signal into first and second
4 subsignals;

first means for combining said first subsignal with a first coset code and for
6 combining said second subsignal with a second coset code orthogonal to said first coset
code so as to produce a first composite coset-encoded signal;

8 means for generating an orthogonal function signal; and

means for modulating said first composite coset-encoded signal with said
10 orthogonal function signal in order to provide a first modulated signal.

2. The transmitter of claim 1 further including:

2 means for generating a pseudorandom noise signal of predetermined PN code,
and

4 means for combining said first modulated signal with said pseudorandom noise
signal of predetermined PN code so as to provide a first output signal.

3. The transmitter of claim 1 further including:

2 means for demultiplexing said information signal into third and fourth
subsignals,

4 second means for combining said third subsignal with a third coset code and for
combining said fourth subsignal with a fourth coset code so as to produce a second
6 composite coset-encoded signal, said first, second third and fourth coset codes being
mutually orthogonal, and

8 means for modulating said second composite coset-encoded signal with said
orthogonal function signal in order to provide a second modulated signal.

4. The transmitter of claim 3, said transmitter further including:

2   means for generating in-phase pseudorandom noise (PN_I) and quadrature phase pseudorandom noise (PN_Q) signals of predetermined PN codes, and

4   means for combining said PN_I signal with said first modulated signal to provide an I output signal, and for combining said PN_Q signal with said second modulated signal

6 to provide a Q output signal.

5. The transmitter of claim 4, said transmitter further including means for

2 modulating in-phase (I) and quadrature phase (Q) carrier signals of a predefined phase relationship with said I and Q output signals, respectively.

6. The transmitter of claim 1 wherein said first means for combining includes:

2   first means for replicating said first subsignal into first and second identical symbol streams,

4   first means for multiplying each of said symbol streams by a coset code coefficient of said first coset code in order to provide first and second intermediate

6 sequences,

      a first multiplexer for combining said first and second intermediate sequences

8 into a first coset-encoded signal,

      second means for replicating said second subsignal into third and fourth identical

10 symbol streams,

      second means for multiplying said third and fourth symbol streams by a coset

12 code coefficient of said second coset code in order to provide third and fourth intermediate sequences,

14   a second multiplexer for combining said third and fourth intermediate sequences into said first coset-encoded signal, and

16   means for combining said first and second coset-encoded signals into said first composite-encoded signal.

7. The transmitter of claim 1 wherein said means for combining said first and

2 second coset-encoded signals includes means for converting said first and second coset-encoded signals into integer values selected from a set of integers including +1

and -1.

8. A transmitter for modulating a set of p information signals of equivalent data
2 rate for simultaneous transmission in a spread spectrum communication system,
comprising:
4 means for combining each of said information signals with one of a set of p
coset codes to produce a set of p coset-encoded signals;
6 means for combining said p coset-encoded signals and for generating a composite
coset-encoded signal;
8 means for generating an orthogonal function signal; and
means for modulating said composite coset-encoded signal with said orthogonal
10 function signal in order to provide a first modulated signal.

9. The transmitter of claim 8, said transmitter further including:
2 means for generating a pseudorandom noise signal of predetermined PN code,
and
4 means for combining said modulated signal with said pseudorandom noise signal
of predetermined PN code so as to provide a first output signal. . .

10. The transmitter of claim 8 wherein said means for combining said
2 information signals with said coset codes includes:
means for replicating a first of said information signals into a set of p identical
4 symbol streams,
means for multiplying each of said symbol streams by one of a set of p coset
6 code coefficients included within a first of said coset codes in order to provide a set of
p intermediate sequences, and
8 a multiplexer for combining said p intermediate sequences into a first of said
coset-encoded signals.

11. The transmitter of claim 5, said transmitter further including means for
2 transmitting said I-modulated and said Q-modulated carrier signals over I and Q
communication channels, respectively.

12. A spread spectrum communication system for modulating an information signal to be transmitted on in-phase (I) and quadrature phase (Q) using a carrier signal and a replica of said carrier signal in phase quadrature therewith, said system including a transmitter comprising:

means for demultiplexing said information signal into first and second sets of subsignals;

means for combining said first set of subsignals with a first set of orthogonal coset codes to produce a first composite coset-encoded signal, and for combining said second set of subsignals with a second set of orthogonal coset codes so as to produce a second composite coset-encoded signal;

means for generating an orthogonal function signal;

means for generating in-phase pseudorandom noise (PN_I) and quadrature phase pseudorandom noise (PN_Q) signals of predetermined PN codes; and

means for combining said PN_I signal with said first composite coset-encoded signal and said orthogonal function signal to provide an I modulation signal, and for combining said PN_Q signal with said second composite coset-encoded signal and said orthogonal function signal to provide a Q modulation signal.

13. The system of claim 12 further including:

means for modulating said carrier signal with said I modulation signal and for modulating said replica of said carrier signal with said Q modulation signal in order to provide I-modulated and Q-modulated carrier signals, respectively, and

means for transmitting said I-modulated and said Q-modulated carrier signals over said I and Q communication channels.

14. The system of claim 13 further including a receiver comprising:

means for producing an estimate of said information signal in accordance with said I-modulated and Q-modulated carrier signals received over said I and Q communication channels.

15. The communication system of claim 14 wherein said receiver further includes means for producing intermediate received signals by demodulating said

received carrier signals using a replica of said orthogonal function signal.

16.   The communication system of claim 15 wherein said receiver further
2   includes:
       means for generating a first despreading signal by replicating said PN_I signal,
4   and
       first means for correlating said intermediate received signals using said first
6   despreading signal in order to provide a first set of in-phase (I) and quadrature phase
   (Q) projection signals.

17.   The communication system of claim 16 further including:
2   means for combining said orthogonal function signal with a pilot signal in order
   to provide a modulated pilot signal,
4   means for transmitting said modulated pilot signal over a pilot channel.

18.   The communication system of claim 17 wherein said receiver further
2   includes:
       means for demodulating said modulated pilot signal transmitted over said pilot
4   channel,
       means for producing an estimate of said pilot signal transmitted over said pilot
6   channel.
       first phase rotation means for generating said estimate of said first information
8   signal on the basis of said first set of said I and Q projections and said estimate of said
   pilot carrier signal.

19.   The communication system of claim 18 wherein said receiver further
2   includes:
       means for generating a second despreading signal by replicating said PN_Q signal,
4   and
       second means for correlating said intermediate received signals using said second
6   despreading signal in order to provide a second set of in-phase (I) and quadrature phase
   (Q) projection signals.

20. The communication system of claim 19 wherein said receiver further
2 includes second phase rotation means for generating an estimate of said second
information signal on the basis of said second set of I and Q projections and said
4 estimate of said transmitted pilot carrier signal.

21. The communication system of claim 18 wherein said receiver further
2 includes means for delaying said first set of I and Q projection signals.

22. A method for modulating an information signal for transmission in a spread
2 spectrum communication system, comprising the steps of:
demultiplexing said information signal into first and second subsignals;
4 combining said first subsignal with a first coset code and combining said second
subsignal with a second coset code orthogonal to said first coset code so as to produce
6 a first composite coset-encoded signal;
generating an orthogonal function signal; and
8 modulating said first composite coset-encoded signal with said orthogonal
function signal in order to provide a first modulated signal.

23. The method of claim 22 further including the steps of:
2 generating a pseudorandom noise signal of predetermined PN code, and
combining said first modulated signal with said pseudorandom noise signal of
4 predetermined PN code so as to provide a first output signal.

24. The method of claim 23 further including the steps of:
2 demultiplexing said information signal into third and fourth subsignals,
combining said third subsignal with a third coset code and combining said fourth
4 subsignal with a fourth coset code so as to produce a second composite coset-encoded
signal, said first, second third and fourth coset codes being mutually orthogonal, and
6 modulating said second composite coset-encoded signal with said orthogonal
function signal in order to provide a second modulated signal.

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25. The method of claim 24 further including the steps of:

2　　　　generating in-phase pseudorandom noise (PN$_I$) and quadrature phase pseudorandom noise (PN$_Q$) signals of predetermined PN codes, and

4　　　　combining said PN$_I$ signal with said first modulated signal to provide an I output signal, and combining said PN$_Q$ signal with said second modulated signal to provide a

6　Q output signal.

26. A method for modulating a set of p information signals of equivalent data

2　rate for simultaneous transmission in a spread spectrum communication system, comprising the steps of:

4　　　　combining each of said information signals with one of a set of p coset codes to produce a set of p coset-encoded signals;

6　　　　combining said p coset-encoded signals and so as to generate a composite coset-encoded signal;

8　　　　modulating said composite coset-encoded signal with an orthogonal function signal in order to provide a first modulated signal.

27. In a code division multiple access (CDMA) communication system, a

2　method for providing in-phase (I) and quadrature phase (Q) spread spectrum communication channels over which is transmitted an information signal, said method

4　comprising the steps of:

　　　　demultiplexing said information signal into first and second sets of subsignals;

6　　　　combining said first set of subsignals with a first set of orthogonal coset codes to produce a first composite coset-encoded signal, and combining said second set of

8　subsignals with a second set of orthogonal coset codes so as to produce a second composite coset-encoded signal;

10　　　　generating in-phase pseudorandom noise (PN$_I$) and quadrature phase pseudorandom noise (PN$_Q$) signals of predetermined PN codes; and

12　　　　combining said PN$_I$ signal with said first composite coset-encoded signal and an orthogonal function signal to provide an I modulation signal, and combining said PN$_Q$

14　signal with said second composite coset-encoded signal and said orthogonal function signal to provide a Q modulation signal.

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28. The method of claim 27 further including the steps of:

2   modulating said carrier signal with said I modulation signal and modulating said

replica of said carrier signal with said Q modulation signal in order to provide

4   I-modulated and Q-modulated carrier signals, respectively, and

          transmitting said I-modulated and said Q-modulated carrier signals over said I
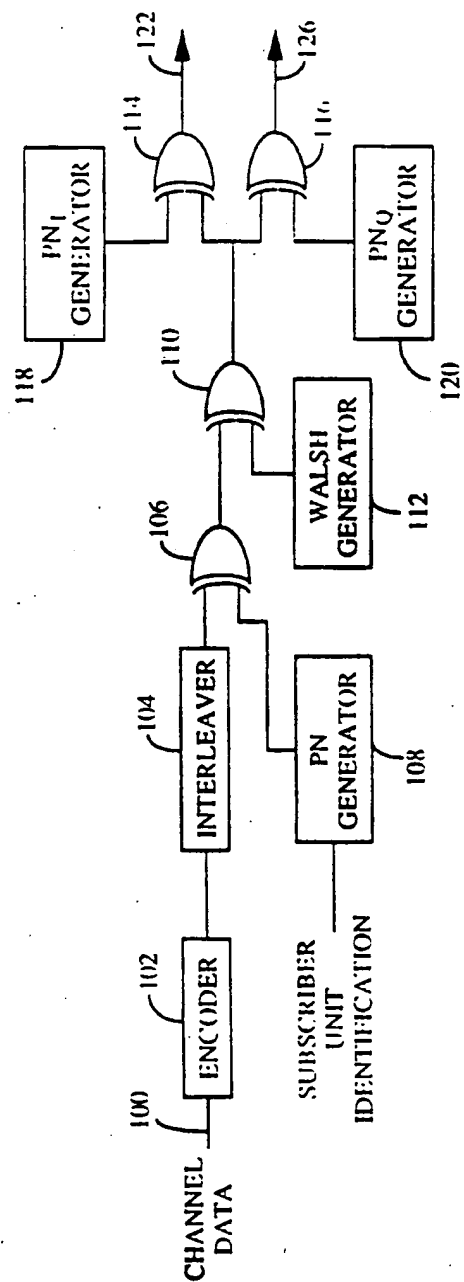
6   and Q communication channels.

29.   The method of claim 28 further including the step of receiving said

2   I-modulated and Q-modulated carrier signals transmitted over said I and Q

communication channels, and for producing an estimate of said information signal in

4   accordance therewith.

30.   The method of claim 29 wherein said step of producing an estimate of said

2   information signal includes the step of demodulating said received carrier signals using

replicas of said orthogonal function signal, said PN_I signal and said PN_Q signal.
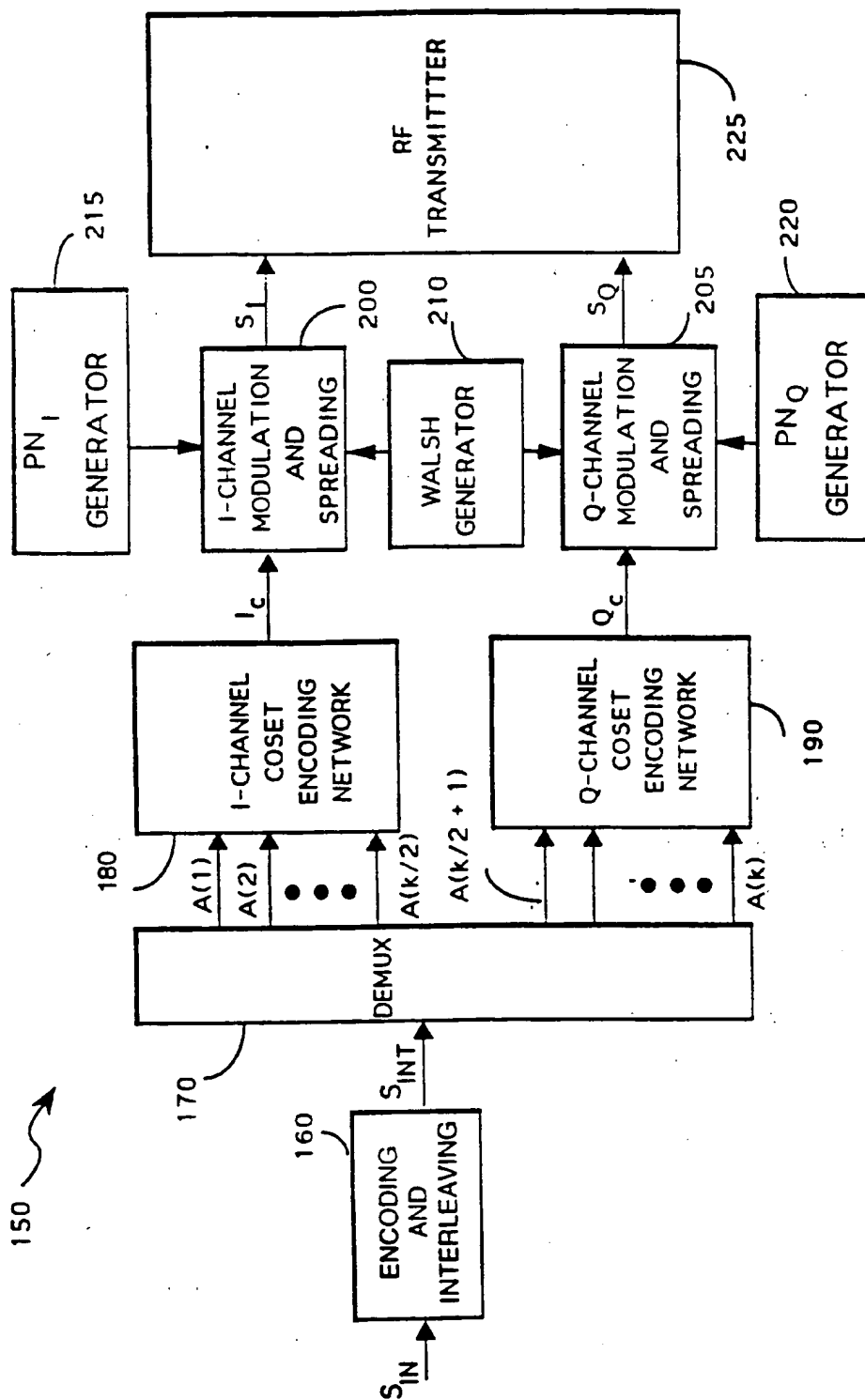
PRIOR ART

FIG. 1

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FIG. 2
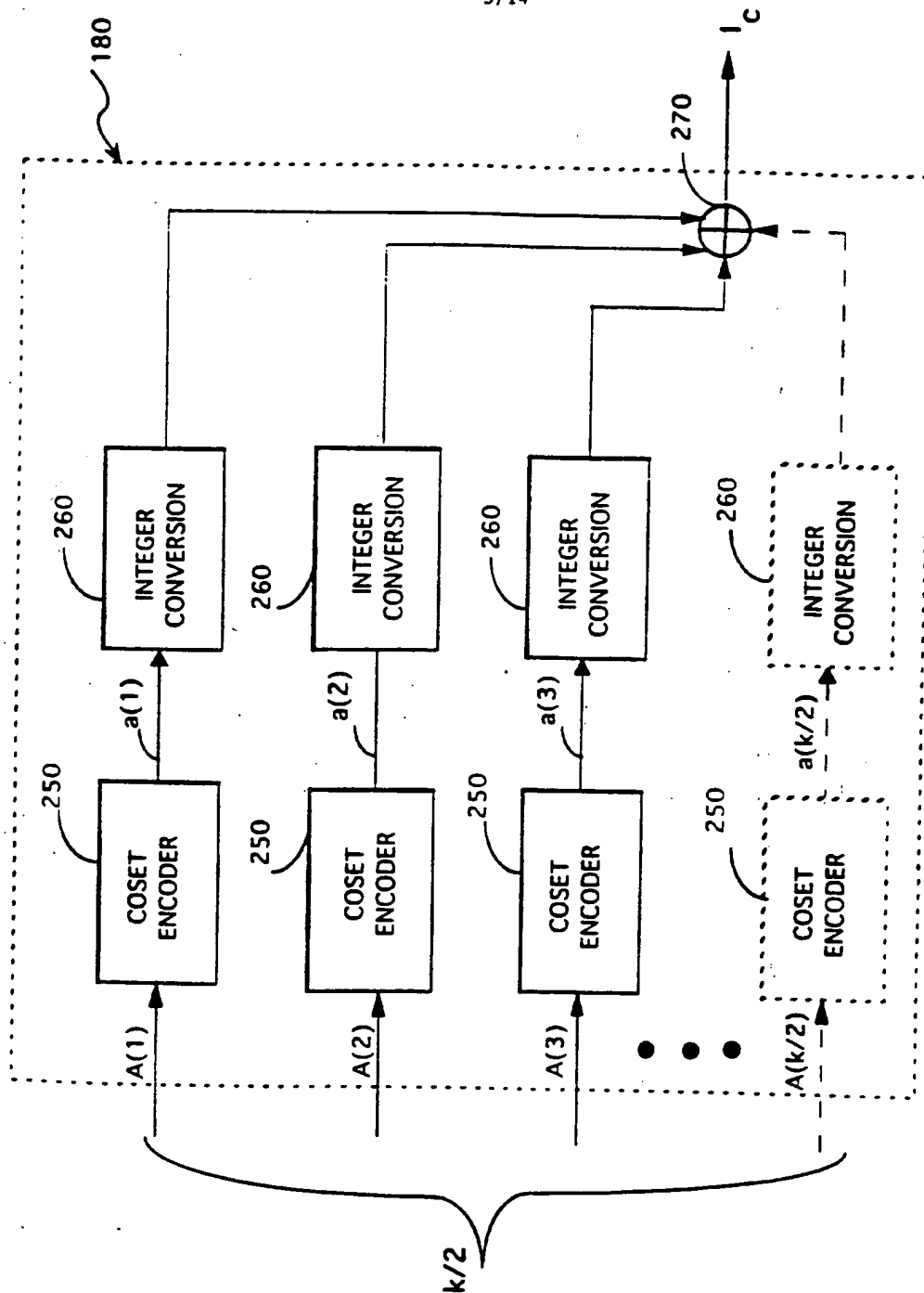
FIG. 3

FIG. 4

FIG. 5

FIG. 6

FIG. 7

FIG. 8

FIG. 9
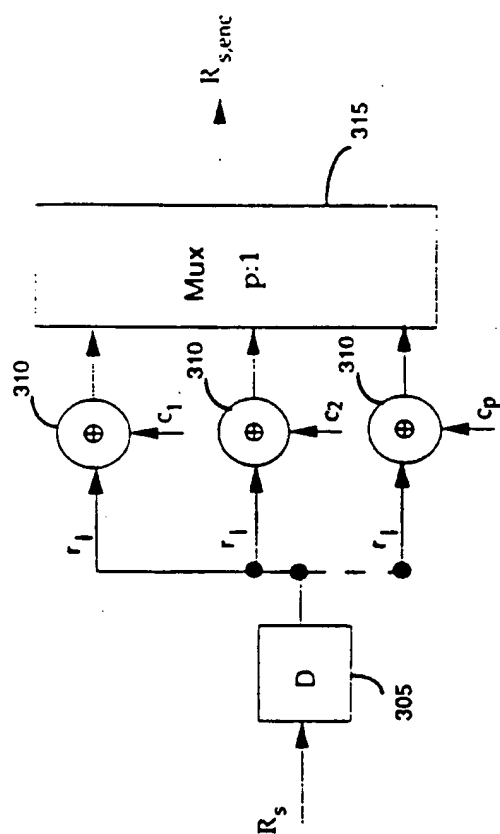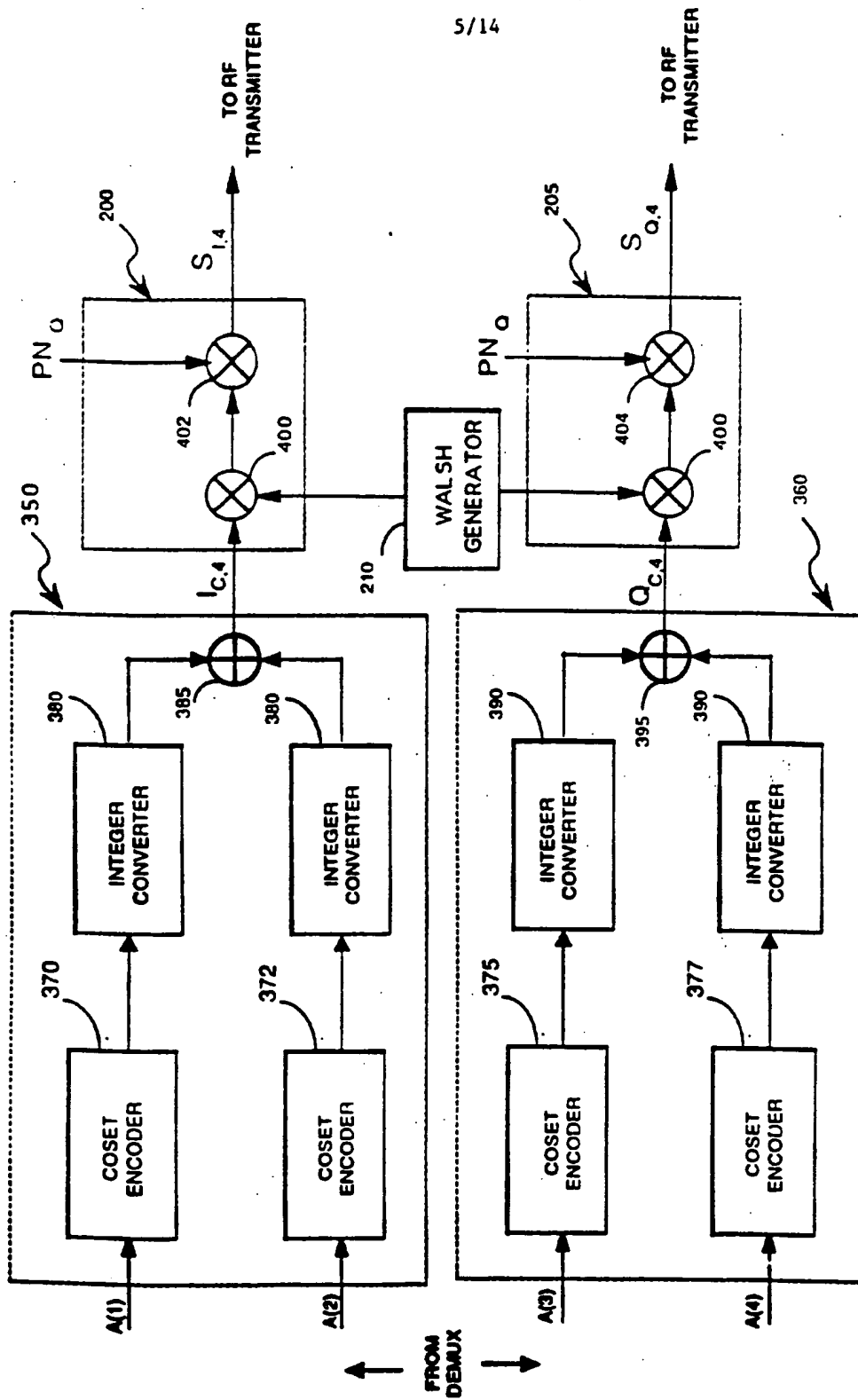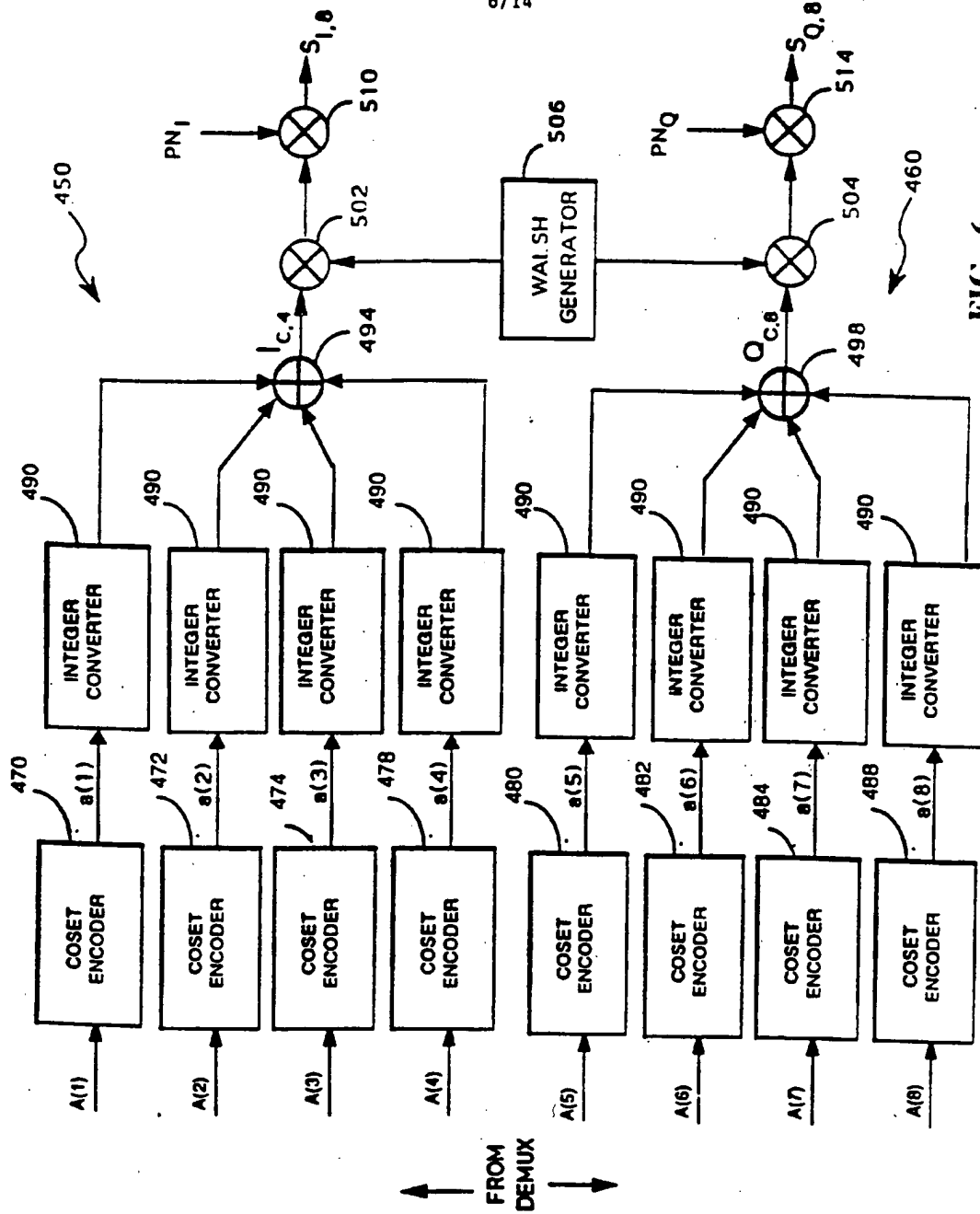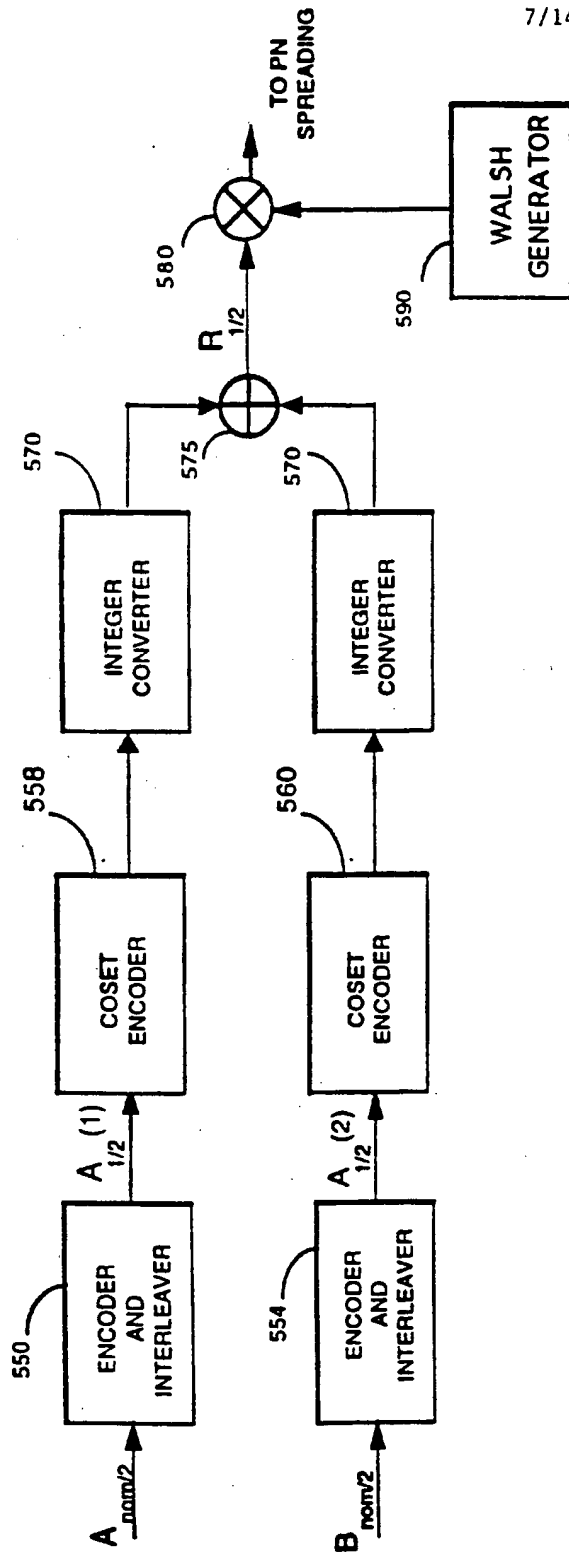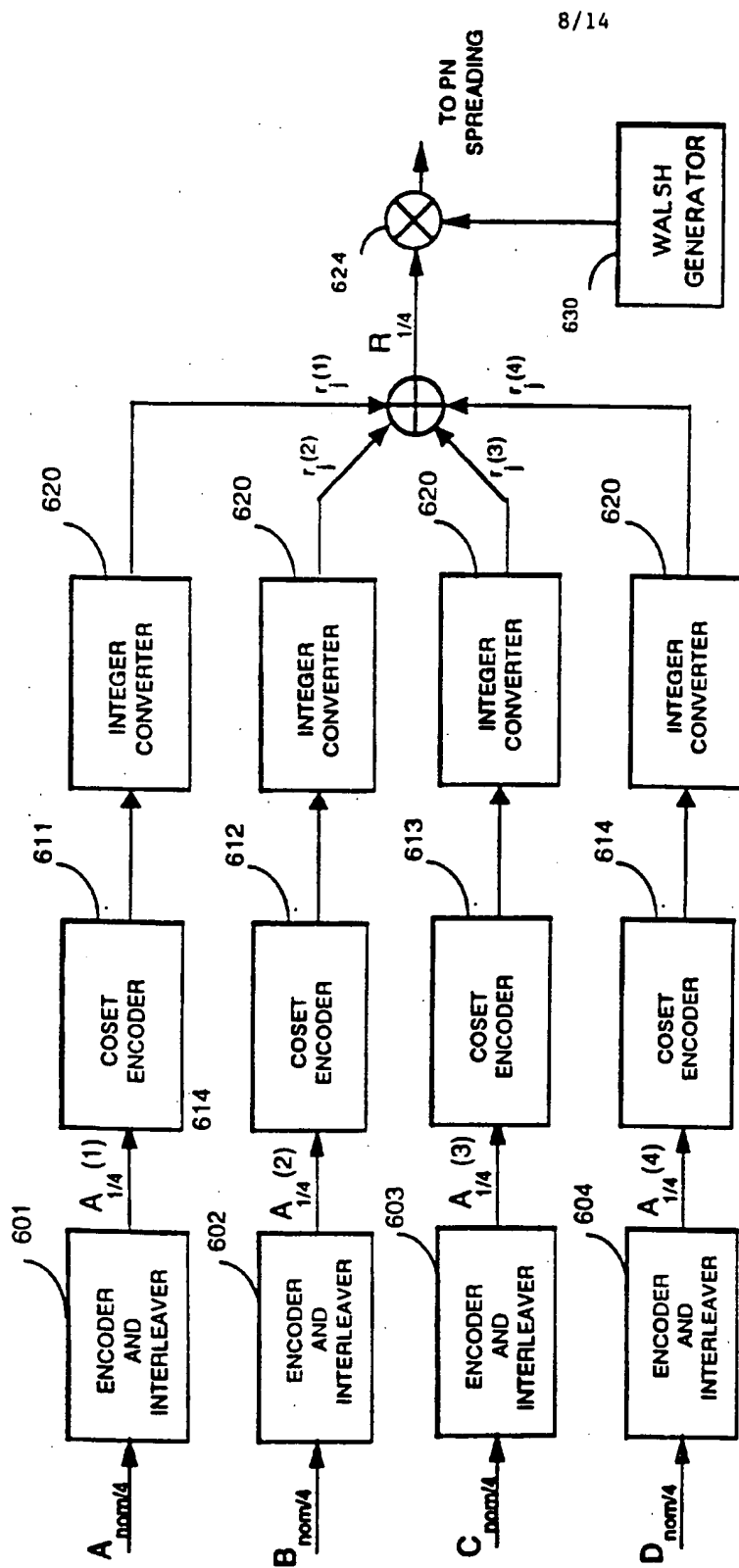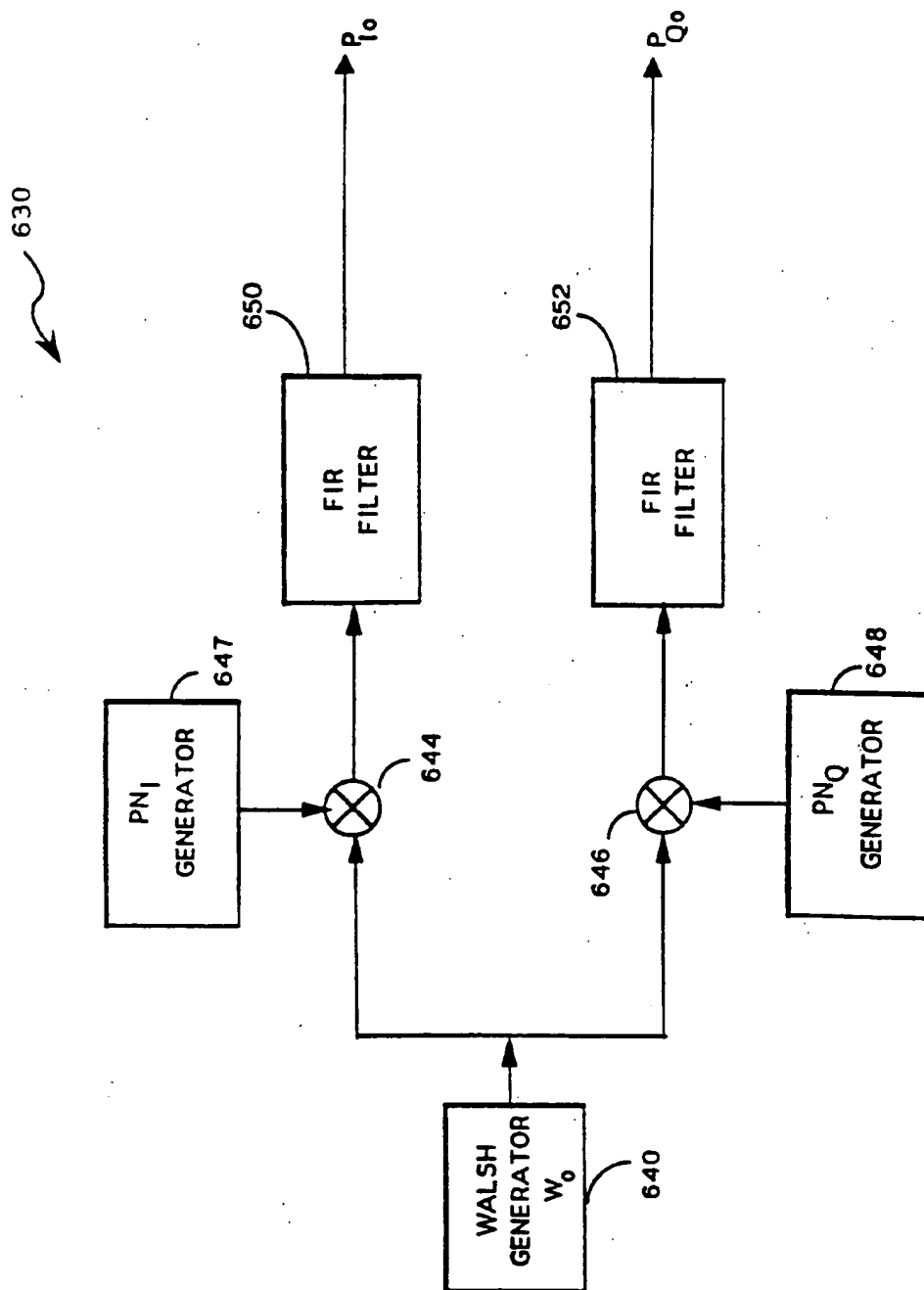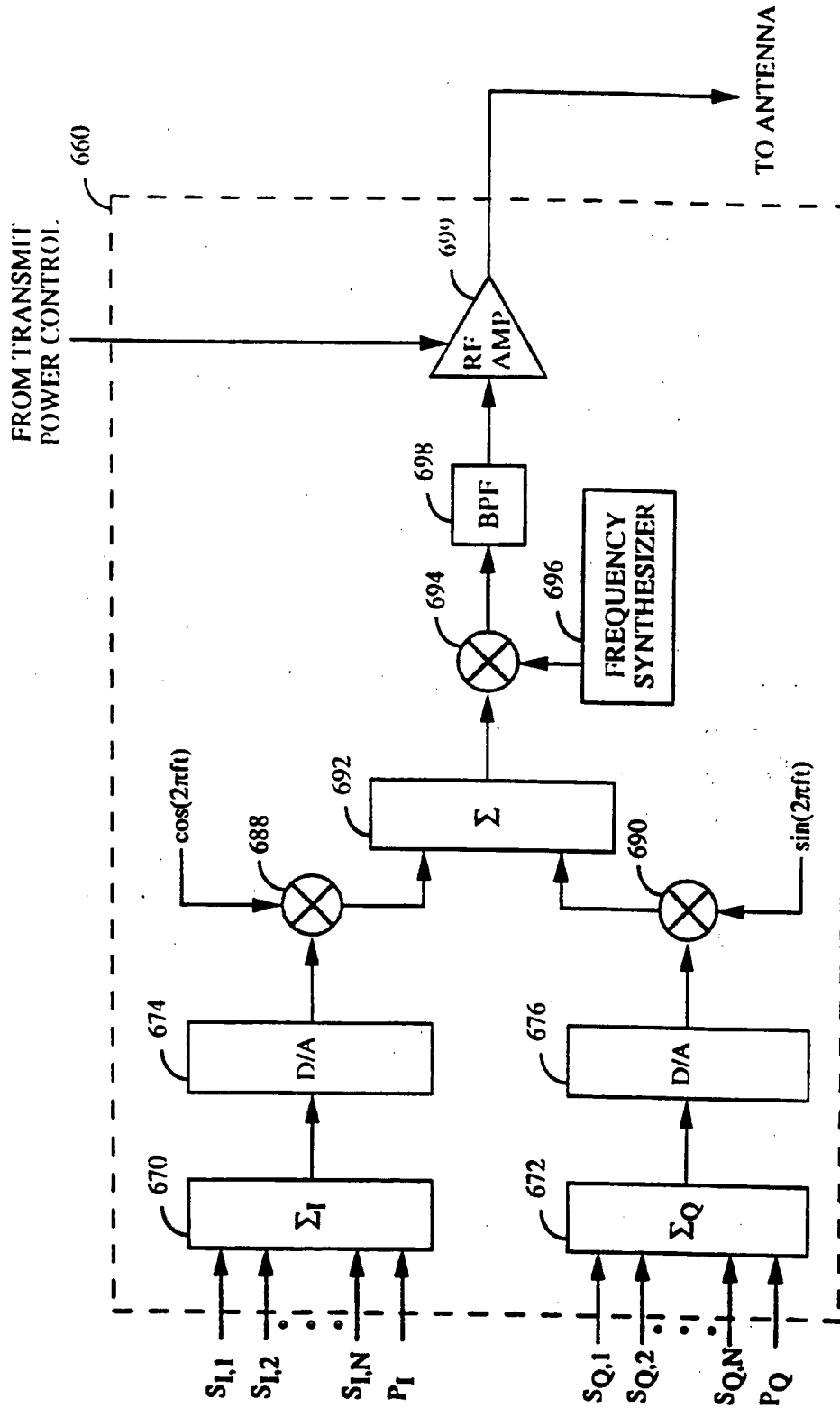
FIG.10
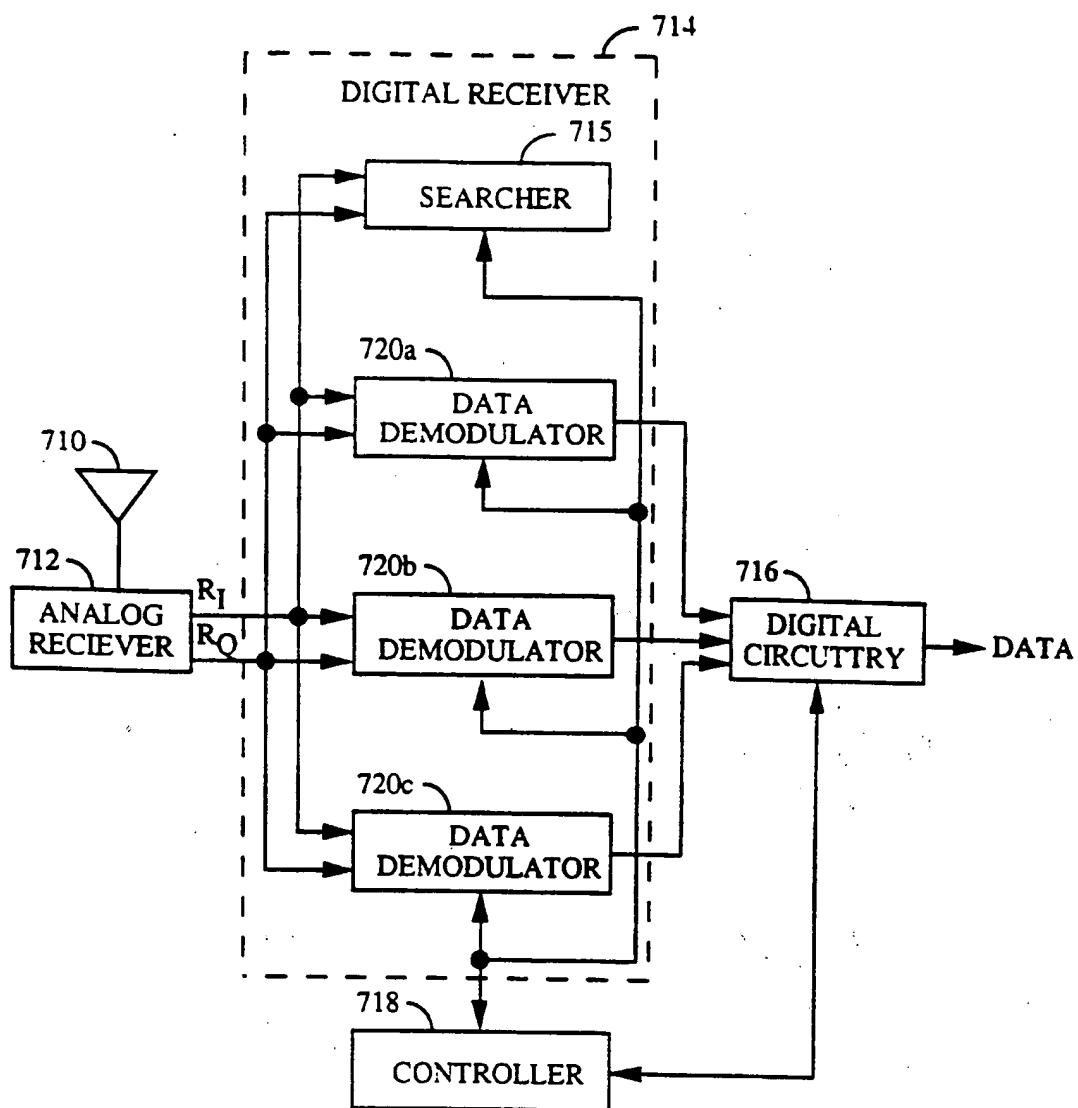
FIG. 11

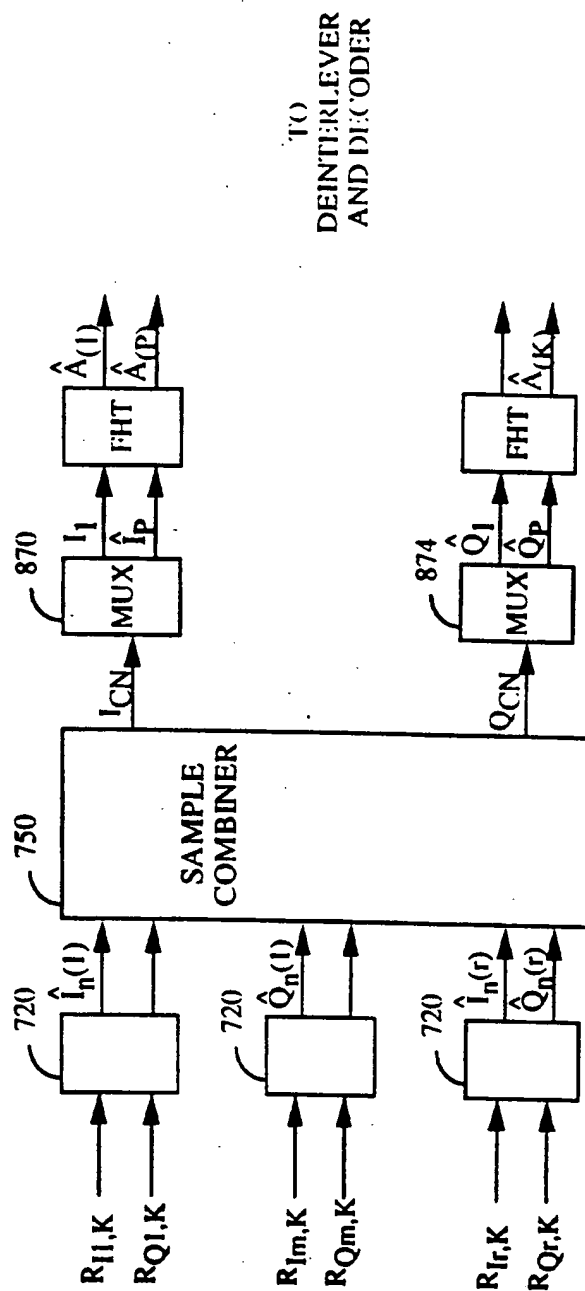**FIG. 12**

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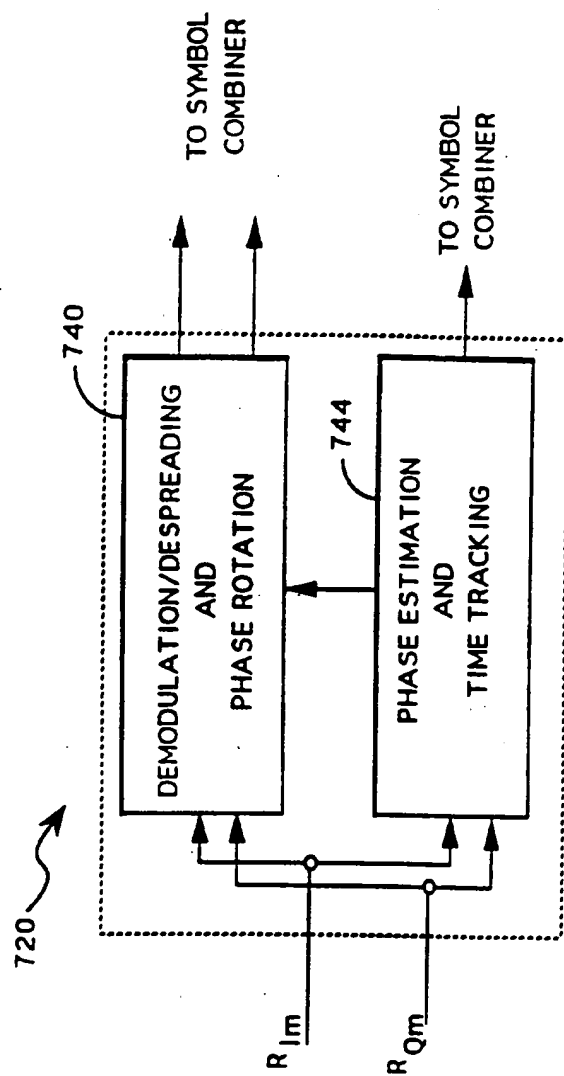**FIG. 13**

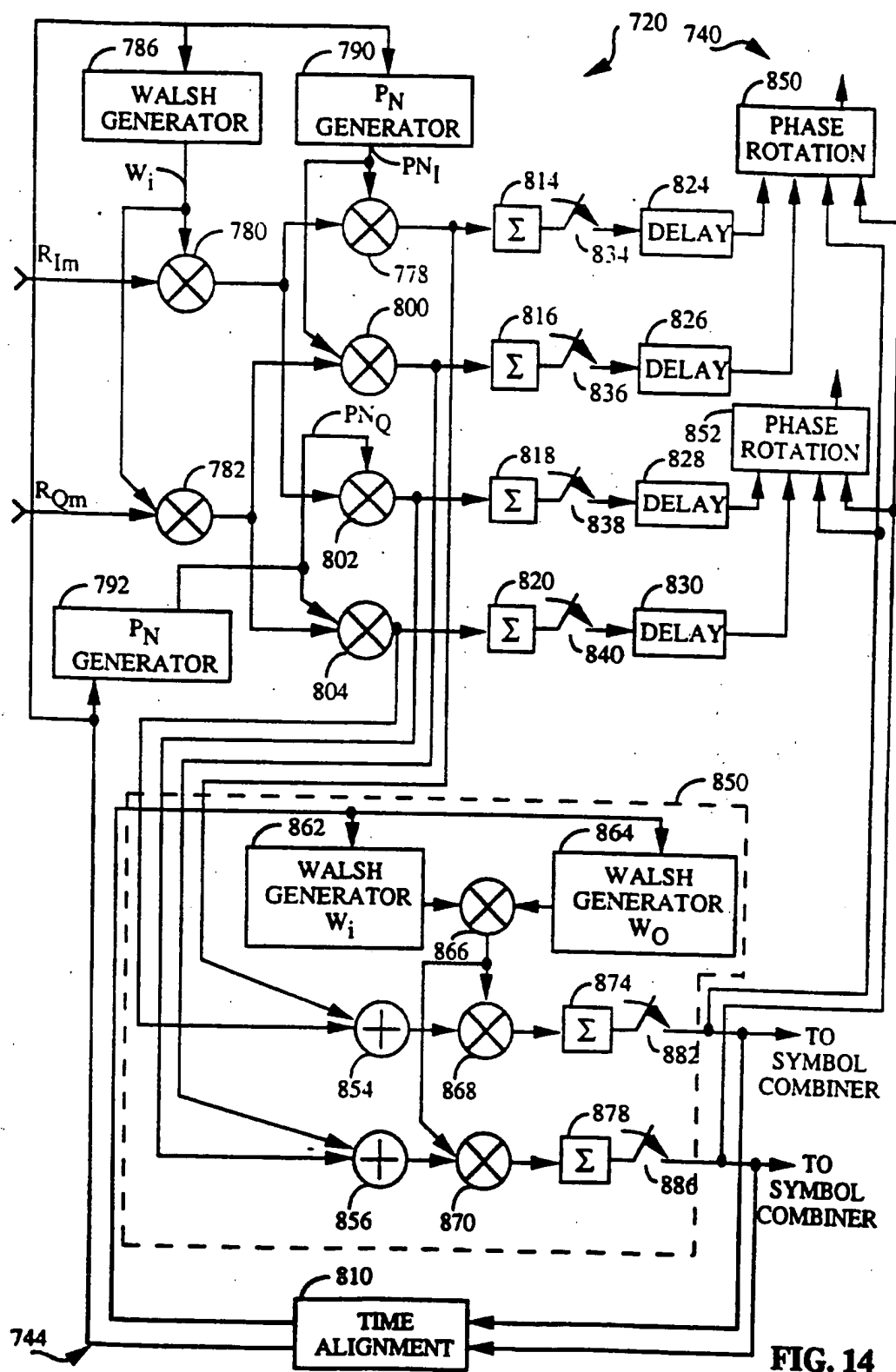**FIG. 14**

INTERNATIONAL SEARCH REPORT

Intu   ional Application No

PCT/US 94/12540

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6   H04J13/00   H04B1/707   H04L27/34

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6   H04B   H04J   H04L   H03M

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	FR,A,2 681 199 (AGENCE SPATIALE EUROPÉENNE) 12 March 1993 see page 7, line 1 - page 8, line 23; figure 1 ---	1,2,4,5, 8,9, 11-13, 22,23, 26-28
A	IEEE TRANSACTIONS ON INFORMATION THEORY, vol.IT34, no.5, September 1988, NEW YORK US pages 1123 - 1151 FORNEY 'Coset codes-Part I: Introduction and geometrical classification' see page 1125, left column, line 30 - right column, line 13; figure 3 ---	

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[X] Further documents are listed in the continuation of box C.   [X] Patent family members are listed in annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
23 February 1995	1 6. 03. 95

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax (+31-70) 340-3016	Bossen, M

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C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	WO,A,92 00639 (QUALCOMM, INC.) 9 January 1992 cited in the application see page 14, line 7 - line 15 see page 25, line 13 - line 20 -----	4,5, 11-13, 25,27,28

Form PCT/ISA/210 (continuation of second sheet) (July 1992)

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		WO-A-	9305589	18-03-93
		EP-A-	0556361	25-08-93
		JP-T-	6502754	24-03-94
		US-A-	5327455	05-07-94
WO-A-9200639	09-01-92	US-A-	5103459	07-04-92
		AU-B-	652956	15-09-94
		AU-A-	8401691	23-01-92
		BG-A-	97222	27-05-94
		CN-A-	1061312	20-05-92
		EP-A-	0536334	14-04-93
		HU-A-	64657	28-01-94
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